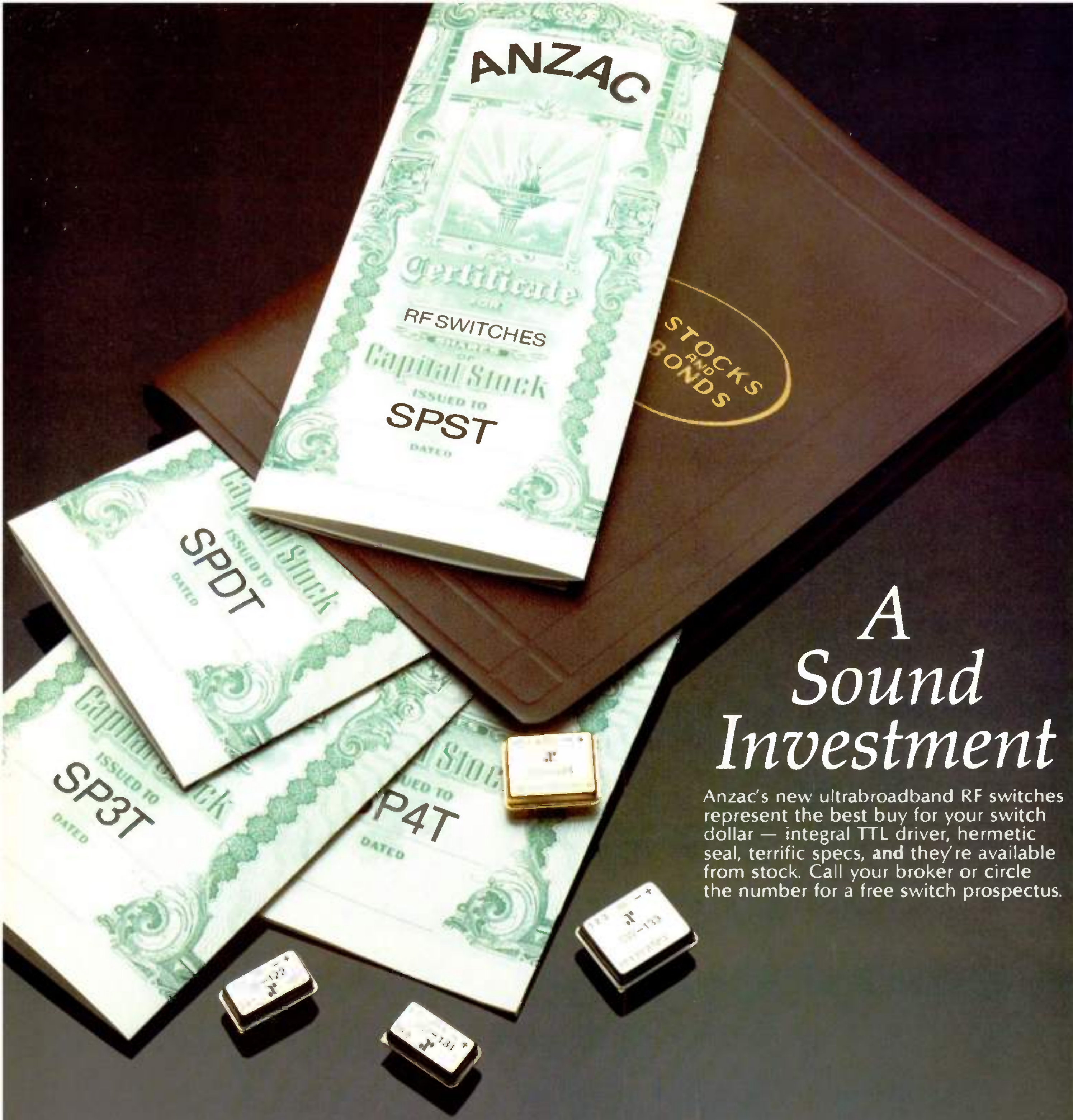
March/April 1983

A Cardiff Publication

rf design

- Phaselock Loops Primer Part I
- Calculator RF Amplifier Design Program
- RF Power Amplifier Testing
- Inductance and Q Measurement



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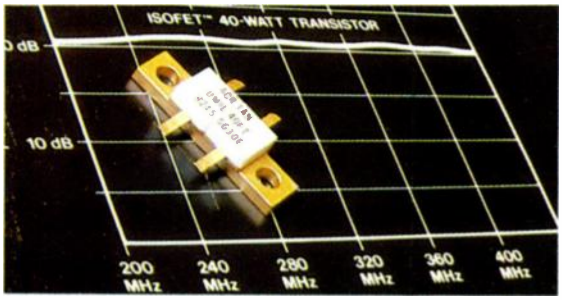
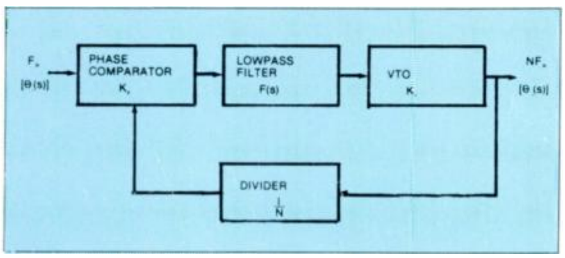
March/April 1983

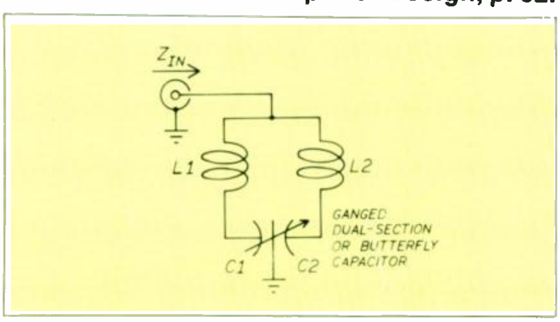


Cover Story, p. 70.



PLL Primer, p. 18.

RF Amplifier Design, p. 32.



RF Power Amp Testing, p. 48.

March/April Cover. The cover, courtesy of Acrian, Inc., announces a new MOSFET fabrication technique enabling power amplifier designers in the VHF and microwave regions to design amplifier stages with higher gain, better isolation and thermal stability. See the cover story on page 70. **1**

PLL Primer: Part I. Taking the mystery out of phase-lock loop design — an integrated approach. **18**

RF Amplifier Design Program. A calculator software package able to evaluate RF amplifier performance and determine required source and load terminations for any condition of stability. **32**

Lumped-Constant Line Stretcher For Testing Power Amplifier Stability. To determine the stability of an RF power amplifier, oscillator, or transmitter under harsh conditions, variable line and fixed attenuator are used to simulate a given VSWR at all phase angles. This article describes a lumped-constant line stretcher using either variable capacitors or inductors. **48**

An Easy Method for Measuring Unloaded Q (Qu) for an Inductor or Filter Tank. A measurement technique to measure inductance by using a narrow-band single pole bandpass filter whose Q approximately equals the unloaded Q of the inductor or tank being measured. **60**

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letters

Dear Sir:

After reading the article "Triple Tuned Circuits" by Andrzej B. Przedpelski in your July/August issue of *r.f. design,* I as a reader was left with the opinion this method of design could be improved in certain areas.

In order to get exact values and symmetrical roll-offs (18dB/octave) on each side of the filter, the following method of design is suggested:

(A) Design Objectives

BW = 40 KHz

Rin = 25 ohms

Rout = 75 ohms

f_1 = −3dB Frequency Low Side

f_2 = −3dB Frequency High Side

$f_c = 36.64\ \text{KHz} = \sqrt{F_1 \times F_2}$

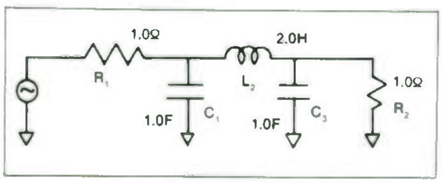
n = 3 (18dB/octave roll-off)

(B) Design a low pass filter, Butterworth type with n = 3, and a cutoff frequency equal to 40 KHz (desired bandwidth) with an input and output impedance equal to 25 ohms.

(C) Using the Bartlett Bisection Theorem and scale the load impedance and associated components to 75 ohms (see appendix).

(D) Complete the filter by resonating (at the center frequency) each shunt capacitor with a shunt inductance and a series capacitor with the series inductor.

Example: Look up the normalized values of a Butterworth Low Pass Filter in any of the following texts on filters: Zveref, Weinberg, White or Geffe. (n = 3)

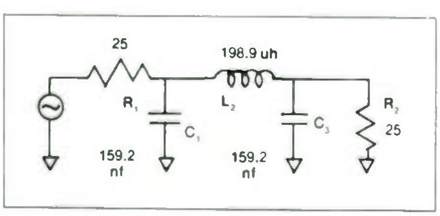


$\omega_c = 2\pi F = 1$ Radian

Scale to 25 ohms (R_1) and 25 ohms (R_2).

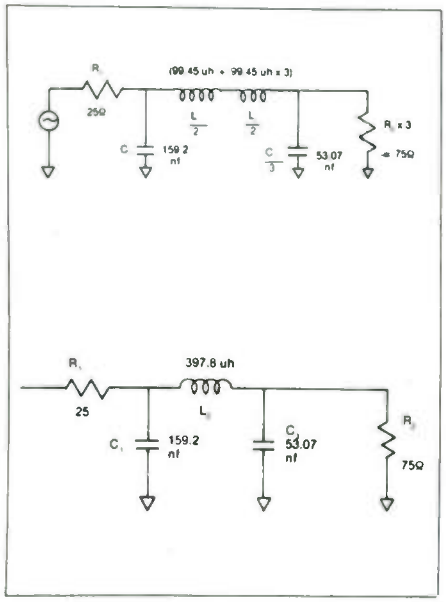
$$C_1 = C_3 = 1/25 \times \frac{1}{2\pi 40E3} \times 1 = 159.2\ \text{nF}$$

$$L_2 = \frac{25}{1} \times \frac{1}{2\pi 40E3} \times 2 = 198.9\ \mu H$$



$\omega_c = 2\pi 40\text{KHz} = 251.3 \times 10^3$ Radians

(E) Use Barlett Bisection Theorem convert load impedance to 75 ohms with associated modified elements.

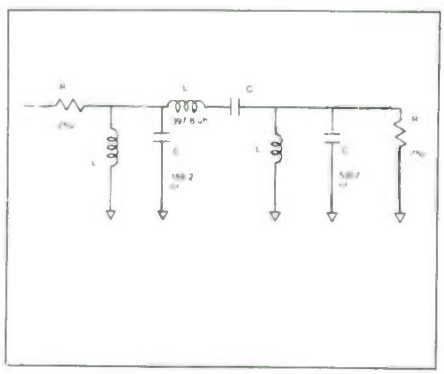


Low Pass Filter, Cutoff = 40 KHz

Rin = 25 ohms

Route = 75 ohms

(F) Convert to Band Pass Filter



Band Pass Filter $f_0 = 34.64$ KHz

$$L_1 = \frac{1}{(2\pi 34.64E3)^2\ 159.2E\text{-}9}$$

$= 132.6\ \mu H$

$$L_3 = \frac{1}{(2\pi 34.64E3)^2\ 53.07E\text{-}9}$$

$= 397.8\ \mu H$

$$C_2 = \frac{1}{(2\pi 34.64E3)^2\ 397.8\ \text{uh}}$$

$= 53.07$ nF

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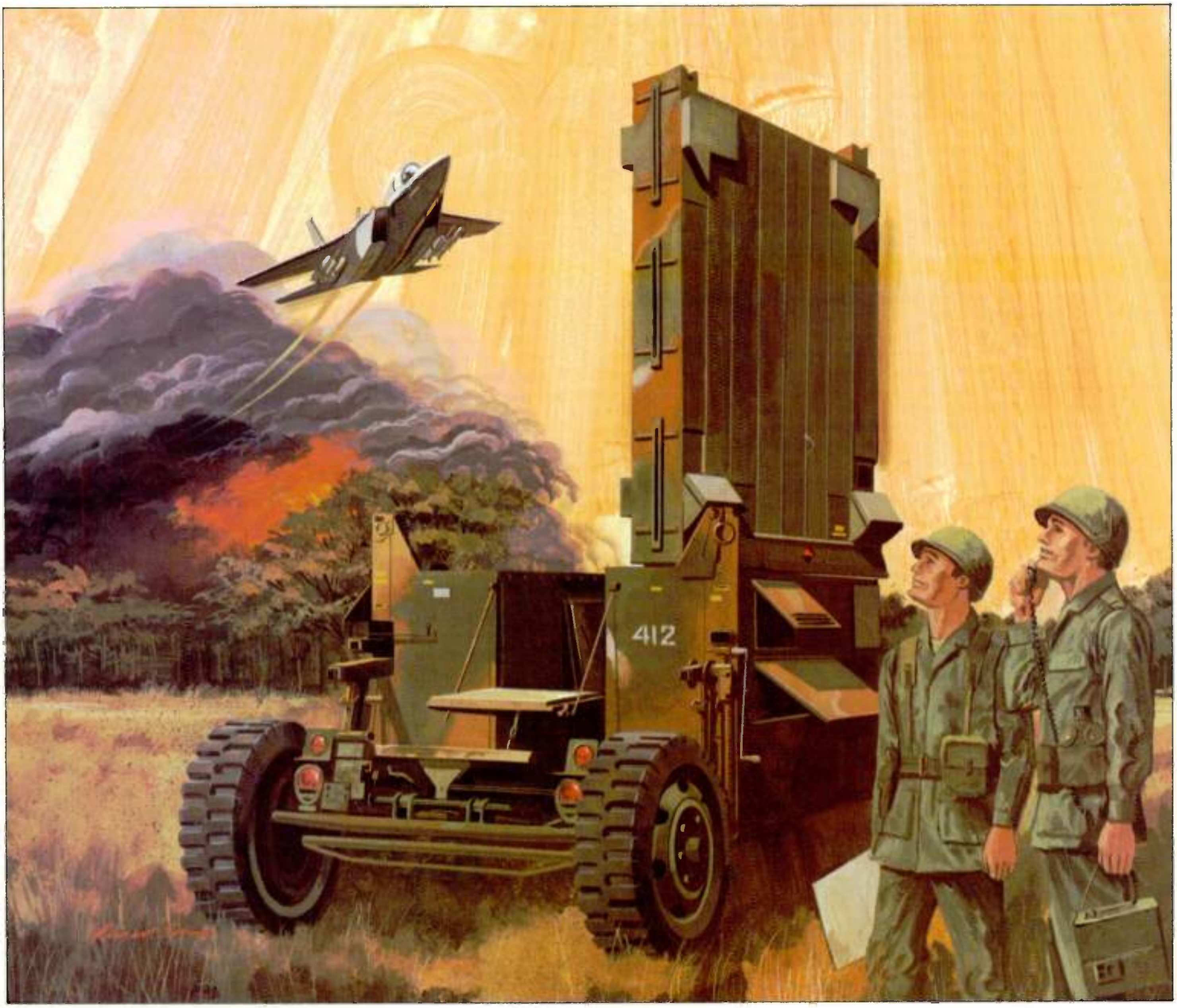

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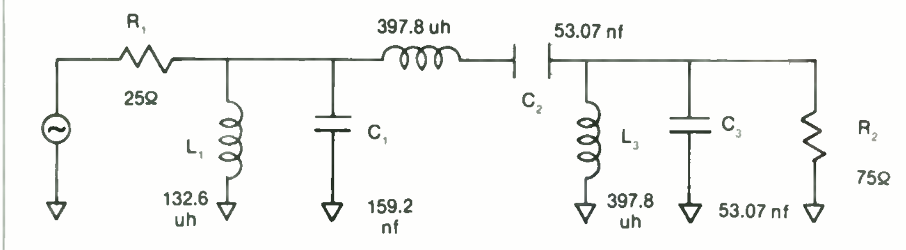


Final Circuit

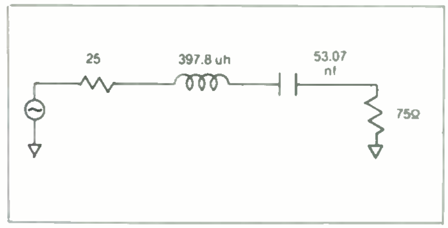
Band Pass Filter BW = 40 KHz
f_0 = 34.36 KHz
Roll-Off = 2πn/octave where n = 3
= 18dB/octave

BW Check (parallel ckt.)

$$BW = \frac{1}{2\pi CR} = \frac{1}{2\pi 159.2E\text{-}9 \times 25} = 40\ KHz$$

$$BW = \frac{1}{2\pi 53.07E\text{-}9 \times 75} = 40\ KHz$$

At Resonance, the Zp of L_1 and C_1, and L_3 and C_3, in parallel, are very large and the circuit becomes



BW check (series circuit)

$$BW = \frac{R}{2\pi L} = \frac{100}{2\pi 397.8\ uh} = 40\ KHz$$

$$Q = \frac{1}{R_s}\sqrt{L/c}$$

$$Q = \frac{1}{100}\sqrt{\frac{397.8\ uh}{53.07\ nf}} = .8658$$

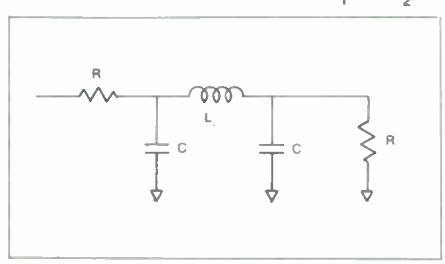
$$BW = \frac{f_0}{Q}$$

$$BW = \frac{34.64E3}{.8658} = 40\ KHz$$

Appendix

Bartlett Bisection Theorem enables an RF designer to transform an equally terminated Low Pass Filter to a Low Pass Filter with unequal terminations.

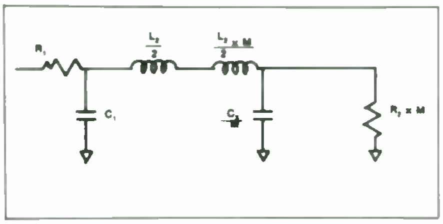
Given Low Pass Filter with $R_1 = R_2$



DETERMINE: Output load resistance desired, to input resistance ration, M

$$\frac{R_2}{R_1} = M \text{ where } R_1 < R_2$$

$M < 10$

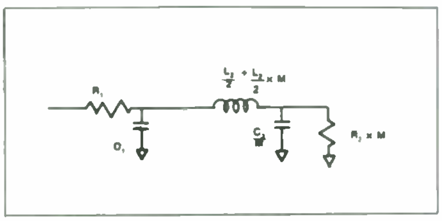


Divide L_2 by two; multiply one-half by M and sum the results —

$$\left(\frac{L_2}{2} + \frac{L_2}{2} \times M\right)$$

Divide C_3 by M
Multiply R_2 by M

Final Filter



I enjoy reading *r.f. design* and find it a very interesting and informative magazine.

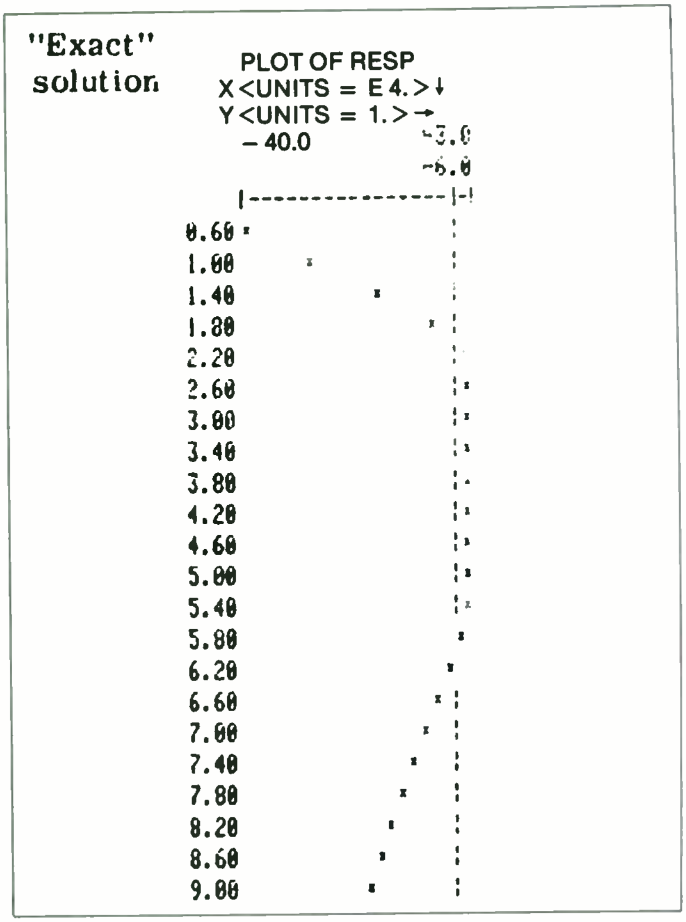
Tony Tortorella

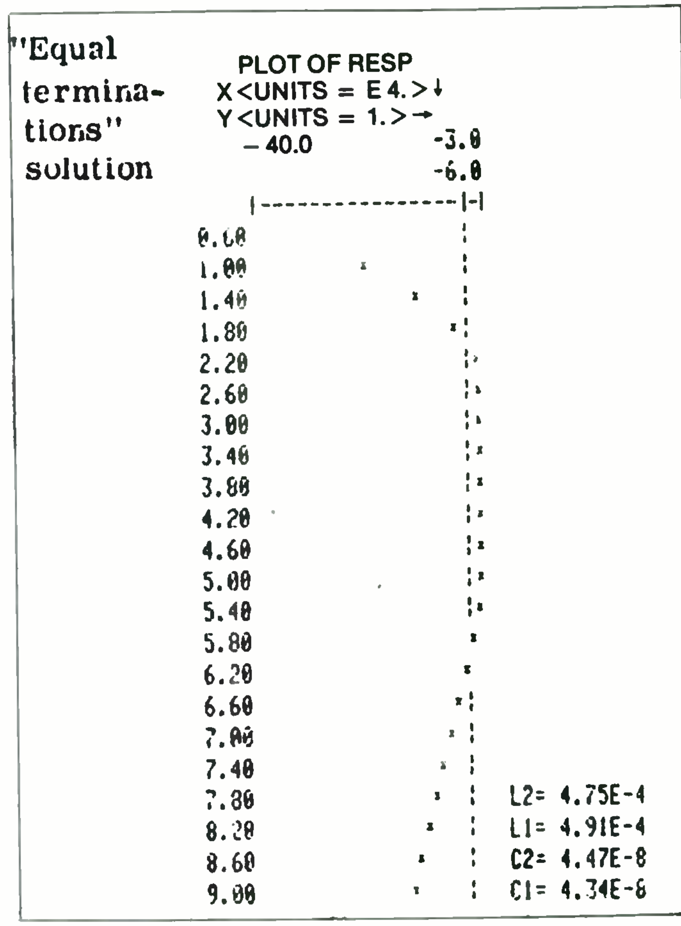
Dear Editor: Tortorella is correct. His solution, even though not completely exact (it neglects the Q of the inductors), is a more elegant one for unequal terminations. The final roll-offs of both filters is 18dB/octave, since both are third order filters. The simple method described in the article was basically meant to be used for equal termination applications and the unequal termination example was used only to show that it can also be used in these cases with minimal sacrifice in performance.

The enclosed figure shows the difference in response between the two filters. The main difference is in the high frequency performance. The equal termination design (when used with unequal source and load) starts rolling off a little slower and does not reach the 18 dB per octave slope until it reaches about 90 KHz.

Using equal terminations both methods give the same filter component values and performance.

A. Przedpelski





Dear Sir:

The November/December 1982 issue of *r.f. design* contained an interesting and informative article by Andrzej B. Przedpelski entitled "Loop Antenna Design." There are several points not brought out in the article which may be of significance in low noise design. This letter discusses these points.

The loop equivalent circuit presented in the various figures contains a resistance R_L. It is never mentioned that this resistance is really the sum of a radiation resistance, R_r, and several components of loss resistance (total loss resistance = R_ℓ). An efficiency factor, n, can be defined for the antenna and is given by

$$n = \frac{R_r}{R_r + R_\ell} = \frac{R_r}{R_L}$$

For many loops, the loss resistance is much higher than the radiation resistance. The primary loss mechanisms are the ohmic loss in the wire, the losses introduced by the ferrite material, and losses caused by proximity effects when multiple turns are used. The efficiency impacts upon the overall system noise figure.

The article also does not mention the effects of atmosphere and man-made noise on received signal to noise ratio. Optimum performance for low noise applications at frequencies below 30 MHz occurs when the system is atmospherically noise limited. The noise figure of a lossless antenna can be defined as

$$f_a = \frac{\text{available noise power rec'd by antenna}}{kTB}$$

(this is, or course, akin to the concept of antenna temperature). The loss occurring in the antenna can be accounted for by introducing a lossy network at the output of the lossless portion of antenna. When the antenna is connected to a receiver, the equivalent circuit is as shown in Figure 1a. The noise figure for this circuit including atmospheric noise is

$$f = f_a - 1 + \frac{f_{rec}}{n}$$

where f_{rec} is the receiver noise figure. For optimum performance in low noise applications,

$$f_a - 1 > f_{rec}/n$$

When this occurs, no further improvement in signal to noise can happen since the system is atmospherically noise limited. It is quite easy, depending upon the electrical size of the antenna, the frequency, the type of active device, etc., for the receiver noise figure (f_{rec}/n) to greatly exceed atmospheric noise. It is therefore important to consider atmospheric noise and antenna efficiency.

Finally, the article does not completely model the noise performance of the antenna/receiver combination. Figure 1b shows an equivalent noise circuit using one of the standard noise models for a two port network. This circuit corresponds to Figure 6 in the article. $\overline{en}^2$ and $\overline{in}^2$ are the equivalent noise sources of the amplifier referred to its input. Z_s is the impedance of the antenna/tuning circuit. The receiver's noise figure is

$$f_{rec} = 1 + \frac{\overline{en}^2 + |Z_S|^2\overline{in}^2 + 2\text{Real}\{Z_S < \overline{en} * \overline{in}_s >\}}{4kTB\ \text{Real}(Z_s)}$$

where $< \overline{en} * \overline{in} >$ is the correlation between $\overline{en}^2$ and $\overline{in}^2$. $\overline{en}^2$ and $\overline{in}^2$ are often frequency dependent. Also depending upon the relative magnitude of $\overline{en}^2$ and $\overline{in}^2$, Z_s, which is also frequency dependent, can have a profound effect on f_{rec}. This is particularly true if

$$\text{Imag}(Z_s) >> \text{Real}(Z_s)$$

or if the real part of Z_s is small.

The points discussed above are of importance in low noise designs where optimum sensitivity is required. They are especially important in wide band designs where f_a can change by large amounts and the various frequency dependencies become significant.

Thank you for your consideration.

Sincerely,
Robert Sainati

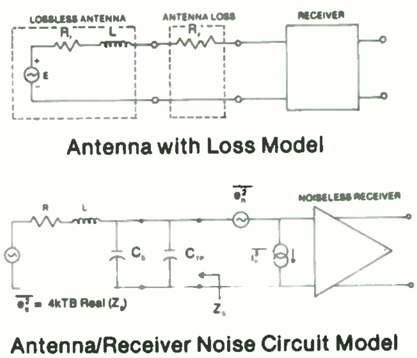


Figure 1. Antenna/Receiver Equivalent Circuits

Dear Editor:

Following are comments on reader Sainati's letter:

Reader Sainati brings up interesting and theoretically valid points. Because of limited space, the article was meant to be a practical design aid since very little information is available in the literature.

1. R_L is defined in Fig. 2 as loss resistance due to coil Q only. Radiation resistance is neglected since as Snelling (Ref. 2) states, "it is usually negligible compared with the losses". This accounts for the well known low efficiency of common ferrite loop antennas (small effective height).

2. Only the method of optimizing the circuit was described, since atmospheric noise is not under the designer's control, and is usually not part of the receiver specifications. It is usually the "system" engineer who determines the equipment requirements and the "design" engineer who has to meet them.

3. The article stressed more the broadband rather than a single frequency design. Thus, the noise density rather than noise figure calculation was treated because of some of the problems described by reader Sainati. This approach has worked satisfactorily in broadband low frequency low noise applications in connection with long range propagation studies.

A. Przedpelski

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Part I

PLL Primer

Taking the mystery out of phaselock loop design — an integrated approach.

By Andrzej B. Przedpelski
A.R.F. Products, Inc.
R & D Laboratory
Boulder, Colorado

Phase lock loops have been with us for decades[1] and yet there seems to be an aura of mystery associated with their design. Actually, their design is fairly straightforward, providing all factors are considered and the design procedure is closely followed. The majority of conventional PLLs (Figure 1) perform one of three functions:

- if N is unity, the PLL is a tracking filter;
- if N is larger than unity, but fixed, the PLL is a frequency multiplier;
- if N is a selectable value, the PLL is a frequency synthesizer.

While the above may be an oversimplification, it is adequate to illustrate the majority of PLL applications.

Type and Order

The first step in PLL design is to select suitable PLL order and type. These are adequately described in a Motorola data book[2]. Briefly, in a conventional PLL, the order and type are determined by the lowpass filter configuration. Two types are in general usage: type 1, using a passive lowpass filter and type 2, using an active filter. Type 2 is usually the more desirable.

Second order PLL is the most commonly described in the literature for the above applications. Even Gardner[3] states, "It is certainly the most prevalent . . . " " . . . second-order loop, as commonly built, is unconditionally stable." However, he also adds: "Parasitic circuit elements often cause a loop intended to be a second order to actually be a much higher order." It is the author's contention that all second order loops are actually of a higher order, as many designers have found out when their "second order" loops oscillated. There always is a pole or two (or more) hidden

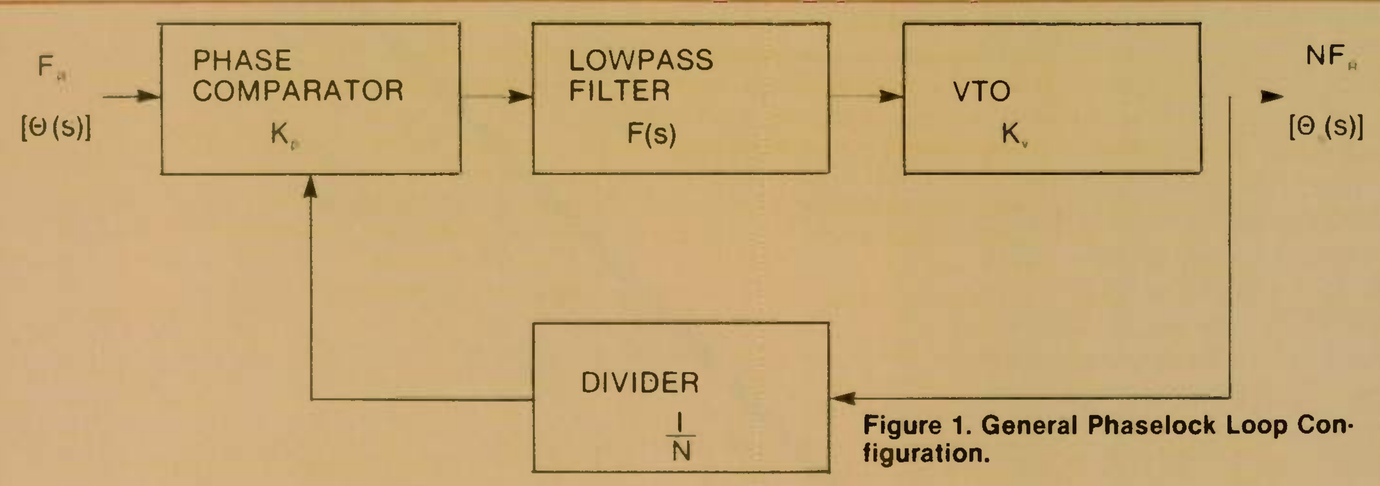


Figure 1. General Phaselock Loop Configuration.

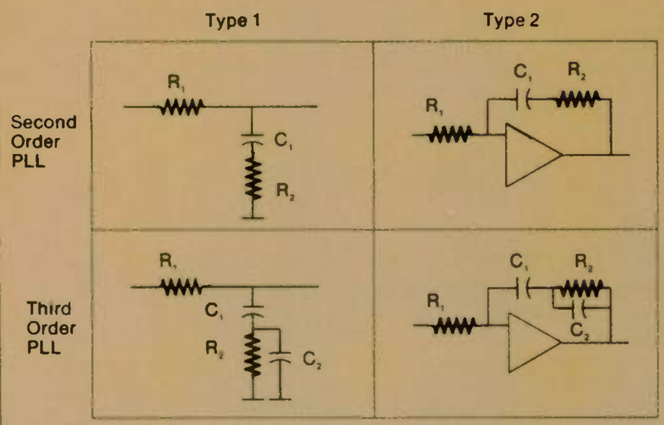


Figure 2. Lowpass Filter Configurations.

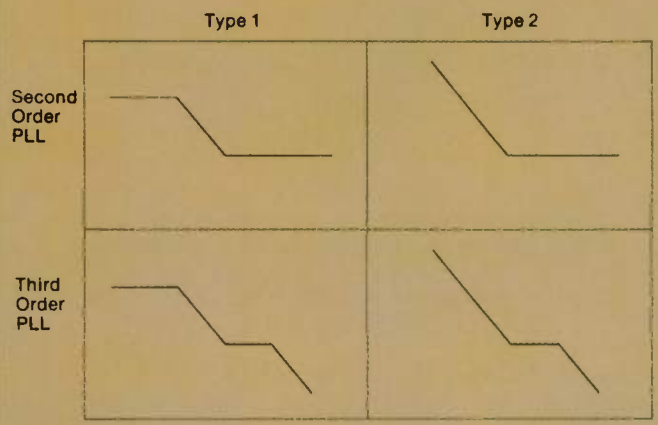


Figure 3. Lowpass Filter Bode Plots for Different PLL Orders and Types.

somewhere. Thus, it is more desirable (and less frustrating in the long run) to design a basically third order loop from scratch, especially since it has advantages over the second order. One important advantage is better suppression of reference frequency sidebands when N is more than one.

The lowpass filter configurations for the different PLL types and orders and their Bode plots are shown in Figures 2 and 3. Because of their inherent advantages, except in some special cases, only third order type 2 PLLs will be considered for further discussion, especially since the first and second orders are well covered in the references. Rohde[4] recognizes the validity of this approach and gives some examples.

Phase Comparator

The next decision is the selection of phase detector type: linear or digital. The most common linear phase comparator is the double-balanced mixer. Except for special applications, such as low noise tracking filters[5], the linear phase comparator is not the best choice. At first glance it seems that it has the advantage of providing a phase error output throughout the signal cycle. However, when N = 1 the divided VTO signal contains phase information in zero crossings only. If the usual programmable divider type is used, this information is only in the leading edge. Another disadvantage is that the gain, K_p, varies with phase error. When no phase error exists (normal steady state operating condition) the gain is maximum and is V/2 volts/radian (Figure 4). It then decreases to zero, when the loop starts skipping cycles. A small advantage of the linear phase comparator is that there is no need to worry about feedback polarity. The loop will choose an operating point where K_p polarity will provide overall negative feedback.

The usual digital phase comparator is actually a frequency/phase comparator, as shown in Figure 4. When

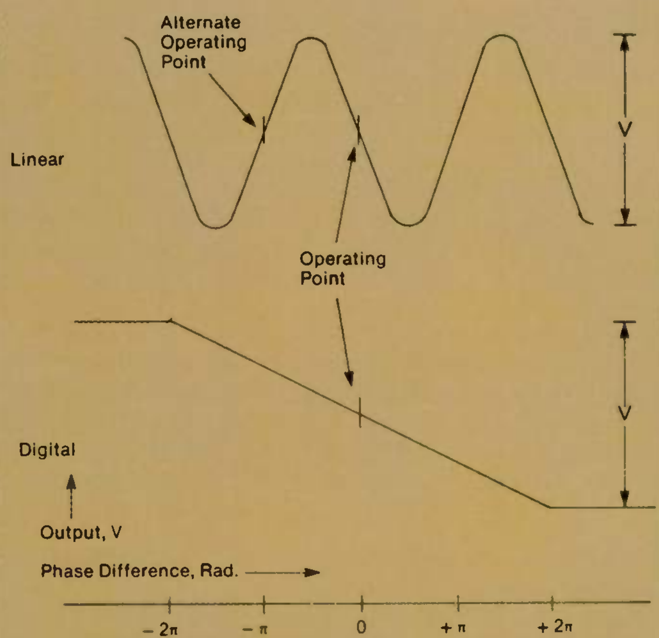
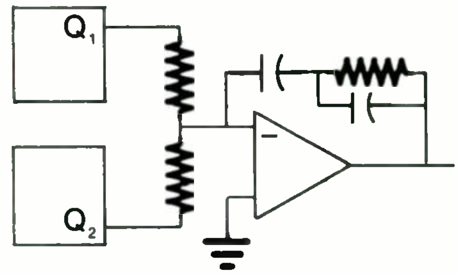
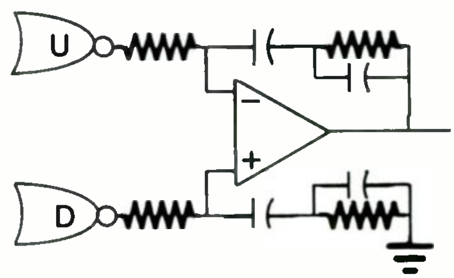


Figure 4. Phase Comparator Characteristics.



Complementary logic (CMOS)



TTL or ECL logic

Figure 5. Digital Comparator/Integrator Interface.

the phase error is larger than 2π it provides a maximum output voltage with a polarity indicating direction of frequency error. This desirable feature, however, makes it necessary to determine the proper K_p, K_v and F(s) polarities to provide an overall negative feedback in the loop. This is no problem if it is done before the PC board is made, since all that is required is to exchange the two comparator inputs, if the original polarity is not correct. The gain, K_p, of the conventional digital comparator is $V/4\pi$ volts/radian. It should be pointed out that, while the curve of Figure 4 is commonly shown as the transfer curve, it is an "effective" and not actual output. The output is actually a series of pulses with a duty cycle which is a function of phase error. Several circuit configurations can be used. If a complementary symmetrical logic is used, such as CMOS, a differential configuration is suitable[6] and the two signals are summed in the integrator (lowpass filter) input; otherwise, the two signals are subtracted in a differential integrator[2]. These two configurations are shown in Figure 5. From these it can be seen that a digital frequency/phase comparator has two outputs which must be either added or subtracted. The digital version's main advantages are constant gain, large pull-in range and fast lock.

A slight glitch may be present in the TTL phase comparator characteristic when the on-chip charge pump is used. This may be undesirable in critical applications. The use of an external integrator reduces this problem considerably and using a faster logic comparator (with the external integrator) practically eliminates it[9].

Integrator

As discussed before, a type 2 PLL uses an active lowpass filter with a pole at zero frequency (ideally) or an integrator. The use of the integrator provides for much better

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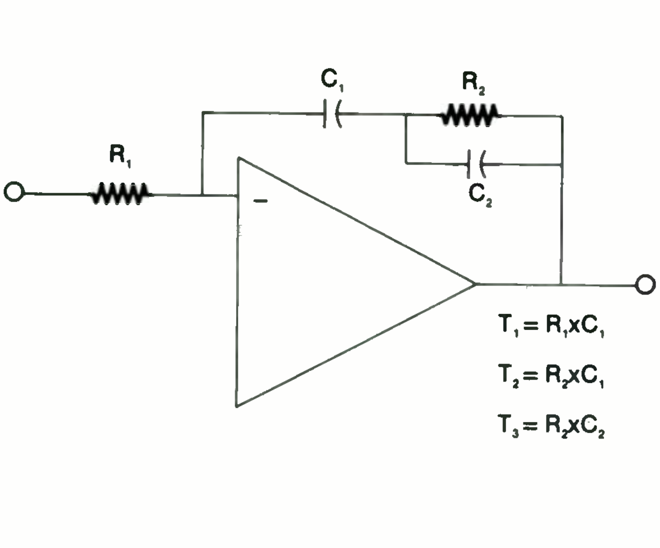

Figure 6. A Third Order Type 2 PLL Integrator.

tracking[2, 3]. A typical circuit is shown in Figure 6. To provide good predictable performance, the integrator response curve should fall below the op-amp open loop response, as shown in Figure 7, down to below one Hz and above T3. The three time constants, T1, T2, and T3 define the integrator.

$$T1 = C1 \times R1 \quad (1)$$
$$T2 = C1 \times R2 \quad (2)$$
$$T3 = C2 \times R2 \quad (3)$$

They can be easily calculated by using the optimizing technique previously described[7]. This technique is much easier

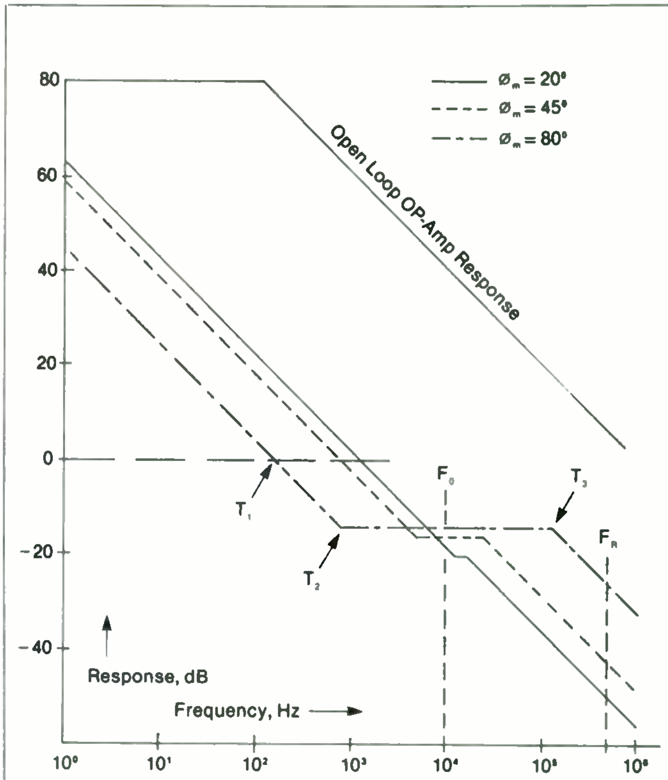

Figure 7. Third Order PLL Integrator Bode Plot.

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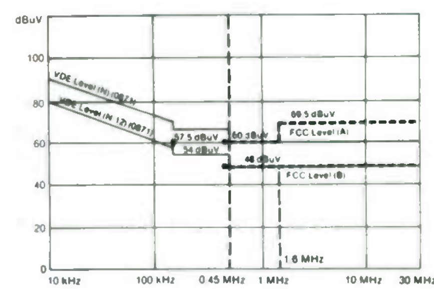

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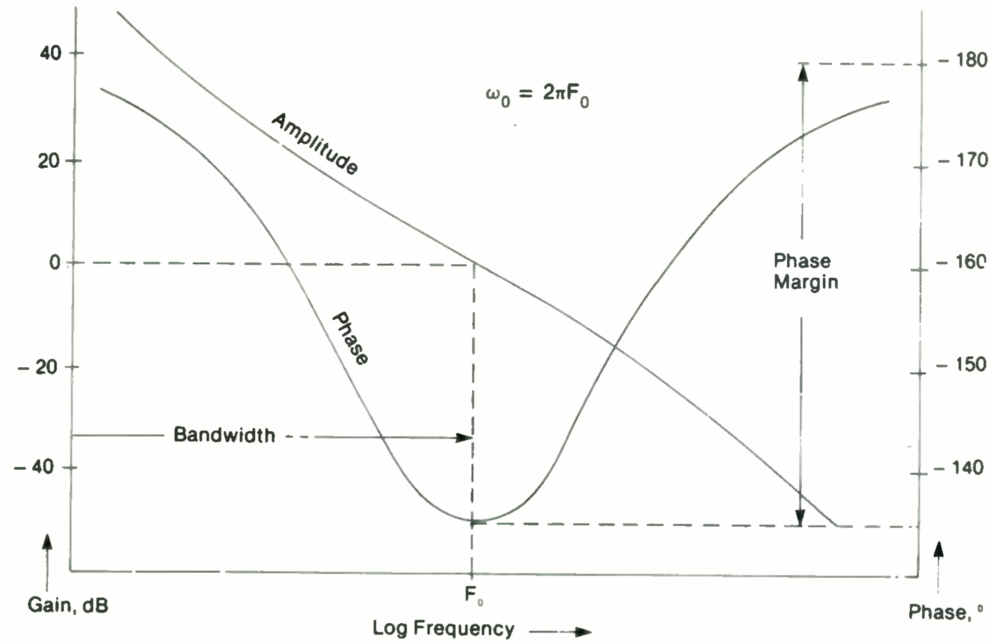

Figure 8. Open Loop Response, G(s)H(s).

to use than the second order approximations[2] usually used and gives more predictable results.

$$T3 = \frac{\sec\phi - \tan\phi}{\omega_0} \quad (4)$$

$$T2 = \frac{2}{\omega_0} \operatorname{Tan}\phi \quad (5)$$

$$T1 = \frac{K_p K_v}{N\omega_0^2}\sqrt{\frac{[\omega_0(T_2 + T_3)]^2 + 1}{(\omega_0 T_3)^2 + 1}} \quad (6)$$

where ϕ = phase margin
and ω_0 = loop "bandwidth"
as defined in Figure 8.

The HP-41 program of Table I simplifies the calculations. In using the three time constants, note that sometimes T2 is defined in the literature as R2(C1 + C2) or T2 + T3, as used in this article. Note also that phase margins over 90° cannot be obtained and that T2 and T3 are independent of K_v, K_p and N. Either a single ended or differential integrator is used depending on the type of phase comparator (Figure 5).

Different bandwidths (such as 3dB, noise, natural frequency) have been used[3] in second order PLL analysis. To facilitate design calculations loop "bandwidth" was defined as 0 to ω_0 radians, as shown in Figure 8. The cut-off frequency, ω_0, is the point at which the open loop gain drops to unity (0 DB). Thus, the phase margin (stability criterion) is the conventional stability margin as used in operational amplifier design[8].

Voltage Tunable Oscillator

Several choices exist, depending on frequency, range and noise characteristics. The easiest to use is the monolithic RC-tuned VTO. This type has very large range and its control characteristic is very linear (K_v is constant). However, its noise characteristics are poor, as seen in Figure 9. The

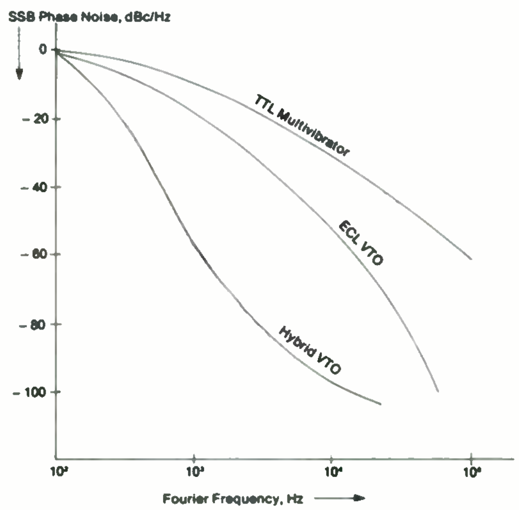

Figure 9. VTO Phase Noise Comparison.

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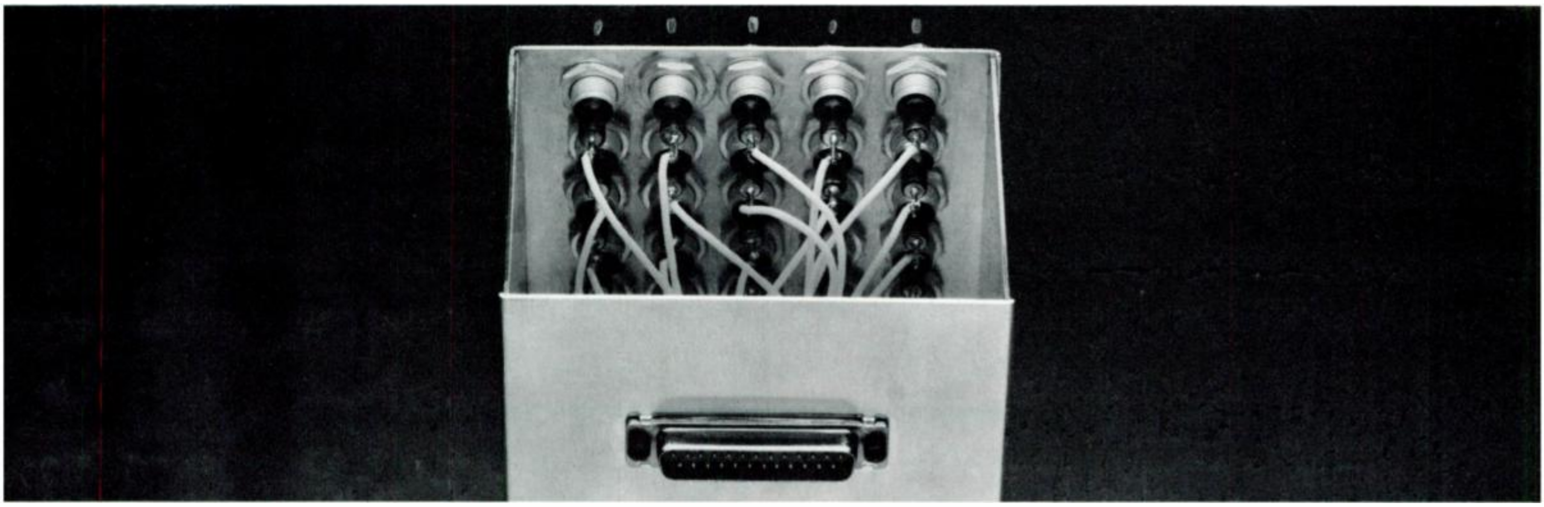

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03 XEQ 01
04 SF 21
05 RCL 19
06 COS
07 1/X
08 RCL 19
09 TAN
10 -
11 RCL 18
12 ST+ X
13 PI
14 *
15 STO 20
16 /
17 STO 14
18 "T3 ="
19 ARCL X
20 AVIEW
21 RCL 19
22 TAN
23 RCL 20
24 /
25 ST+ X
26 STO 13
27 "T2 ="
28 ARCL X
29 AVIEW
30 RCL 14
31 +
32 RCL 20
33 *
34 X↑2
35 1
36 +
37 RCL 14
38 RCL 20
39 *
40 X↑2
41 1
42 +
43 /
44 SQRT
45 RCL 15
46 *
47 RCL 16
48 *
49 RCL 17
50 /
51 RCL 20
52 X↑2
53 /
54 STO 12
55 "T1 ="
56 ARCL X
57 AVIEW
58 RTN
59♦LBL 01
60 "KP = "
61 ARCL 15
62 AVIEW
63 "KV = "
64 ARCL 16
65 AVIEW
66 "N = "
67 ARCL 17
68 AVIEW
69 "FO = "
70 ARCL 18
71 AVIEW
72 "∠ = "
73 ARCL 19
74 AVIEW
75 ADV
76 END
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REGISTERS

R01	
R02	
R03	
R04	
R05	
R06	
R07	
R08	
R09	
R10	
R11	
R12	[T1]
R13	[T2]
R14	[T3]
R15	K_p
R16	K_v
R17	N
R18	Fo
R19	øm
R20	[ω_0]
R21	
R22	
R23	
R24	

	Linearity	Noise	Range	Max. Operating Frequency
RC VTO	very good	very high	very large	low
LC ECL	medium	medium	medium	medium
LC Hybrid	medium	low	medium	high
VCXO	poor	very low	very low	low

Figure 10. VTO Comparison.

LC-tuned ECL VTO has smaller range, lower noise and poorer linearity. The LC-tuned hybrid or discrete VTO has the lowest noise, but again, lower linearity and smaller tuning range.

In some special low noise, low range applications, a crystal controlled VTO may be used[10]. A comparison of these types is summarized in Figure 10.

Frequency Divider

With the exception of the 1:1 tracking loop, some form of frequency division in the feedback path of the loop is needed to provide the required closed loop frequency multiplication. Usually a fixed or programmable digital divider is used. If a programmable divider with a large divide-by-N is used, it is sometimes necessary to use a pulse stretcher to operate the usually low-speed phase comparator.

Part II of this article to appear in the May/June issue will continue the discussion with design parameters and possible problems. Ed.

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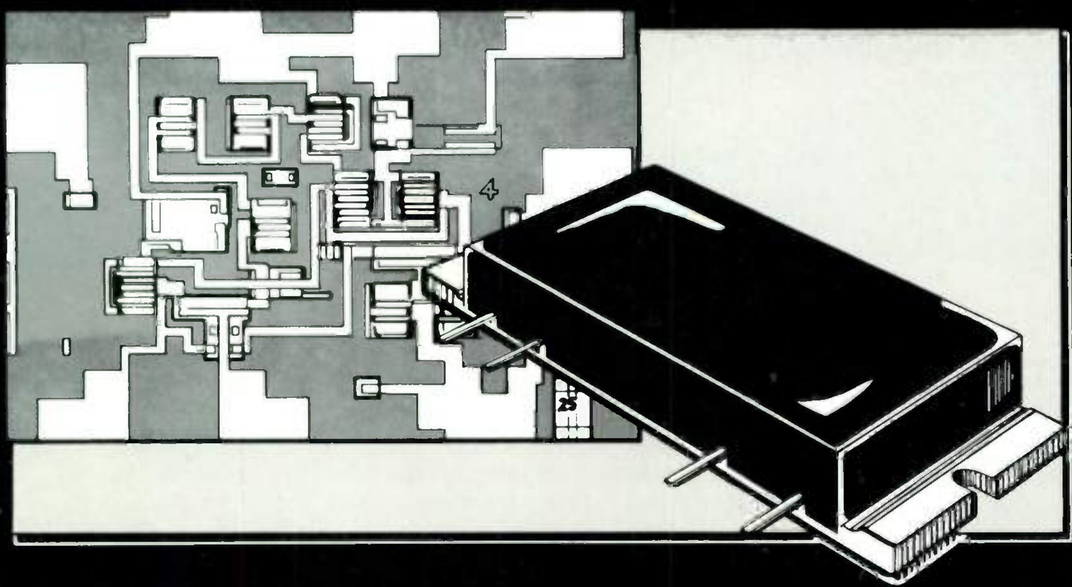

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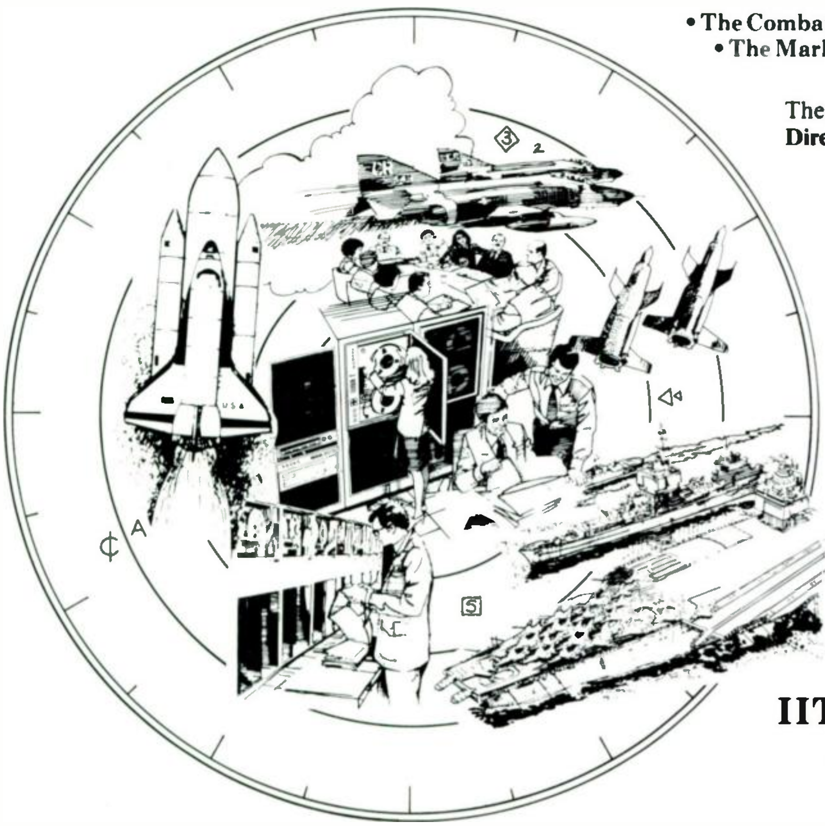

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By R.K. Feeney
D.R. Hertling
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The HP-41 calculator is a very flexible and powerful instrument. Its ability to handle conveniently alphanumeric data makes the construction of highly interactive, easy-to-use programs possible. This article presents a package of RF design software using the aforementioned capabilities of the HP-41. This program will design for all conditions of transistor stability and can utilize either y- or S-parameters.

General Description of the Software

This software gives the designer the ability to determine stability factors, amplifier gains, input and output admittances, and terminations for a specified stability factor. A conversion routine is included so that the design may be accomplished with S-parameters if desired.

The ability of the HP-41 to easily work with alphanumeric data is employed to make all programs interactive. The "customized keyboard" feature is not used, but instead the programs are operated from a "menu" as often done with computers. As a consequence of this technique almost no instructions are needed to use the software. This eliminates many data input errors and is especially time saving if the design programs are used only occasionally.

The software is organized around a main program and eight subroutines. The main program contains the menu and calls appropriate subroutines according to the task selected. Flags keep track of quantities that have been entered or generated, e.g., y-parameters or load terminations. Thus, when a given function is selected, the main program calls only for quantities that have not been entered. If all data are available, the desired calculation is completed.

Upon initial execution of the main program, the calculator prints the menu of possible commands. These are:

1 - LSF
2 - OPT TERM
3 - GAINS
4 - YIN/OUT
5 - CHNG YS/L
6 - CHNG YPAR
7 - K-STERN
8 - YS/L FOR K
9 - S/Y CONV

Commands are executed by entering the appropriate number and pressing R/S. The calculator will prompt the user for any necessary data. Upon execution of any command, the program returns to the command mode displaying the message, CMND. At this time it is ready to execute the function. All entered and calculated data are available for the next operation. The operation of the nine commands will now be summarized.

1. LSF. The first command calculates the Linvill stability factor of the device. It also checks to see if $Re(y_i) < 0$ and $Re(y_o) < 0$. If either real part is negative, the device is potentially unstable and the Linvill factor is not a valid indicator. A message indicating this fact is displayed.

2. OPT TERM. This command calculates the optimum terminations (Y_S and Y_L) and the optimum gain for an in-

herently stable transistor. The routine first checks the Linvill stability factor and prints out an appropriate message if the optimum terminations do not exist.

3. GAINS. The third command calculates the available, operating and transducer power gains. If a particular gain does not exist, a message indicating that fact is given.

4. YIN/OUT. This command causes the input and output admittances to be calculated. It requires the y-parameters and the source and load admittances. The user is prompted to enter these data if they are not already present.

5. CHNG YS/L. Command number five allows the user to change the values of the source and load admittances. Some routines generate Y_S and Y_L, and some require Y_S and Y_L. This routine enables the current values of Y_S and Y_L to be changed without disturbing other data in the calculator.

6. CHNG YPAR. This routine is similar to number five, but it is used to change the current values of the y-parameters.

7.K-STERN. This command calculates Stern's stability factor for the circuit. It needs the device parameters and the source and load terminations.

8. YS/L FOR K. Command number eight calculates the required source and load terminations for a specified value of Stern's stability factor. An iterative procedure is used which stops when the change between successive iterations in the imaginary parts of both Y_S and Y_L is less than or equal to 0.01 percent. Only the device parameters are needed for this calculation.

9. S/Y CONV. The final routine converts S-parameters to y-parameters. Normalized 50-ohm S-parameters are assumed, which are entered in polar form. The program calculates the y-parameters and loads them into the appropriate storage locations. Once the S-parameters have been entered, any of the various commands may be executed.

Preliminary Operations

Before the program can be utilized, it must be loaded into the calculator. The complete program occupies 1821 bytes of code and requires a minimum SIZE of 043. With a SIZE of 043, there will be 17 registers free when all routines are loaded. If S-parameters will not be used the SPAR routine and associated storage will not be needed. In this case only 1512 bytes are used and the minimum SIZE of 030 will leave 74 registers unused. None of the routines use registers 00 and 01 so these are free for temporary storage while using the software. Ten (10) complete magnetic cards are required to store the entire package.

Using the Software

Once loaded, the only other factor to be considered is the presence of the HP82143A printer. The software works best with the printer, but it is completely usable without the device. If the printer is not present, it is recommended that flag no. 21 be set manually before using the programs. With flag no. 21 set, the calculator will stop whenever a print or prompt command is executed. Pressing R/S will continue the program. If flag no. 21 is not set, the calculator rapidly goes through the explanatory statements, but stops on all prompts. If the printer is used, it should be set to the NORMAL mode.

Examples of Program Use

The utility and flexibility of the software will be illustrated by two examples. The first example uses y-parameters and an inherently stable transistor while the second considers a transistor described by S-parameters that is not inherently stable. All features of the programs are illustrated by these examples.

Example I.

This example illustrates several aspects of the software by focusing on an inherently stable transistor specified by y-parameters. The method of design will be discussed in detail with a printout of all entered and generated data listed on the right of the discussion.

Given a 2N5109 transistor operated at 200 MHz. The y-parameters in mmho for the transistor are:

$y_{ie} = 22 + j9 \qquad y_{re} = 0 - j2.2$

$y_{fe} = 40 - j185 \qquad y_{oe} = 1 + j8$

Start the design procedure by executing program AMP. The program displays the command table and requests a command number. Enter 1 to evaluate the Linvill stability factor. The y-parameters are requested. Enter them as requested, pushing R/S each time. The Linvill stability factor is computed and printed. Since it is less than one, the transistor is inherently stable.

After completing the stability factor calculation, the calculator requests the next command. Enter 2 to calculate the optimum load and source admittances. The optimum termination routine always calculates the Linvill stability factor in addition to the optimum terminations and gain. Knowing this, the first command could have been omitted.

Any desired command can now be executed. For example, what are the input and output admittances? Run command No. 4. The result confirms that a conjugate match exists at both ports of an optimally terminated, inherently stable device.

Suppose that the transistor will not be optimally terminated, but that

$$Y_S = 20 - j50 \text{ mmho}$$

and

$$Y_L = 1.5 + j5 \text{ mmho}$$

What will be the various gains under these conditions?

It is first necessary to change the current values of Y_S and Y_L. (Remember, the optimum values of Y_S and Y_L are still stored in the calculator.) To change Y_S and Y_L, execute command no. 5. After entering the new values of Y_S and Y_L, execute routine no. 3 which will calculate all the power gains. Since all the gains are different, neither of the ports is conjugately matched.

The calculator is ready to receive the next command, but it isn't going to get one. This completes Example I.

Example II.

The second example will be done with S-parameters to illustrate how they are handled by the program. In addition, the device used is potentially unstable, so other aspects of the software are demonstrated.

Given a 2N6680 GaAs FET operated at 4 GHz. The 50-ohm S-parameters are:

$S11 = 0.792\angle -107.4 \qquad S12 = 0.070\angle 26$

$S21 = 1.995\angle 81.1 \qquad S22 = 0.659\angle -60.8$

A. Assume that it is desired to design an amplifier having at least 12 dB transducer power gain with adequate stability.

Start the design procedure by executing the design program AMP. The program displays the command table and requests a command.

Since this design will be done with S-parameters, command no. 9 should be entered. The program requests the S-parameters which will be converted to y-parameters and a flag set indicating that the y-parameters are present.

Guess that the transistor is inherently stable. Enter

```
            XEQ "AMP"
ENTR CMND
1-LSF
2-OPT TERM
3-GAINS
4-YIN/OUT
5-CHNG YS/L
6-CHNG YPAR
7-K-STERN
8-YS/L FOR K
9-S/Y CONV

CMND?
            1.     RUN

ENTER Y-PARAMS

GI=?
         22-03     RUN
BI=?
          9-03     RUN

GR=?
             0     RUN
BR=?
       -2.2-03     RUN
CMND?
            2.     RUN

OPTIMUM VALUES

L. S. F. = 923.3E-3

GS= 86.61E-3
BS= -53.00E-3

GL= 3.937E-3
BL= -10.00E-3

GMAX= 17.59E0 dB
GF=?
         40-03     RUN
BF=?
       -135-03     RUN

GO=?
          1-03     RUN
BO=?
          8-03     RUN
L. S. F. = 923.3E-3

CMND?
            4.     RUN
GIN= 86.61E-3
BIN= 53.00E-3

GOUT= 3.937E-3
BOUT= 10.00E-3

CMND?
            5.     RUN
ENTER YS AND YL
GS=?
         20-03     RUN

BS=?
        -50-03     RUN
GL=?
        1.5-03     RUN
BL=?
          5-03     RUN

CMND?
            3.     RUN

OPERATING GAIN
GP= 9.509E0 dB

AVAILABLE GAIN
GA= 16.27E0 dB

TRANSDUCER GAIN
GT= 4.952E0 dB

CMND?
```

Example 1.

```
            XEQ "AMP"
ENTR CMND
1-LSF
2-OPT TERM
3-GAINS
4-YIN/OUT
5-CHNG YS/L
6-CHNG YPAR
7-K-STERN
8-YS/L FOR K
9-S/Y CONV

CMND?
            9.     RUN
MS11=?
          .792     RUN
AS11=?
        -107.4     RUN
MS12=?
          .070     RUN
AS12=?
            26     RUN
MS21=?
         1.995     RUN
AS21=?
          81.1     RUN
MS22=?
          .659     RUN
AS22=?
         -60.8     RUN

CMND?
            2.     RUN

OPTIMUM VALUES

L. S. F. = 1.268E0

DO NOT EXIST


CMND?
            8.     RUN
K=?
            10     RUN

GS= 12.60E-3
BS= -29.75E-3

GL= 10.04E-3
BL= -14.74E-3


CMND?
            3.     RUN

OPERATING GAIN
GP= 13.74E0 dB

AVAILABLE GAIN
GA= 13.74E0 dB

TRANSDUCER GAIN
GT= 12.77E0 dB

CMND?
            5.     RUN

ENTER YS AND YL
GS=?
      10.22-03     RUN
BS=?
     -20.23-03     RUN
GL=?
          1-03     RUN
BL=?
          1-03     RUN

CMND?
            4.     RUN
GIN= 10.58E-3
BIN= 23.65E-3

GOUT= 6.749E-3
BOUT= 14.29E-3


CMND?
            5.     RUN

ENTER YS AND YL
GS=?
      10.22-03     RUN
BS=?
     -20.23-03     RUN
GL=?
      6.749-03     RUN
BL=?
     -14.29-03     RUN

CMND?
            7.     RUN

K-STERN= 6.457E0

CMND?
            3.     RUN

OPERATING GAIN
GP= 14.48E0 dB

AVAILABLE GAIN
GA= 11.85E0 dB

TRANSDUCER GAIN
GT= 11.85E0 dB

CMND?
```

Example 2.

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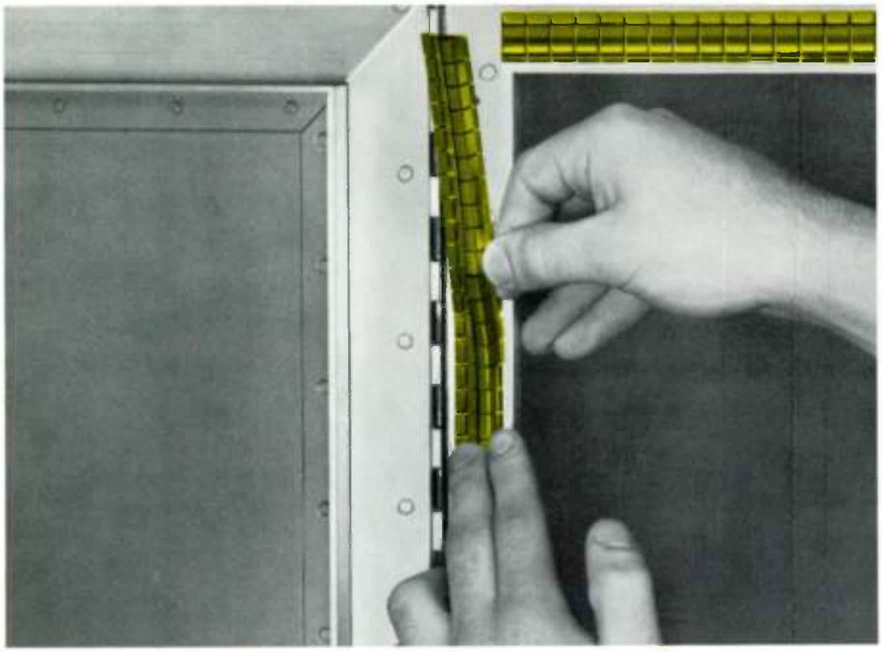

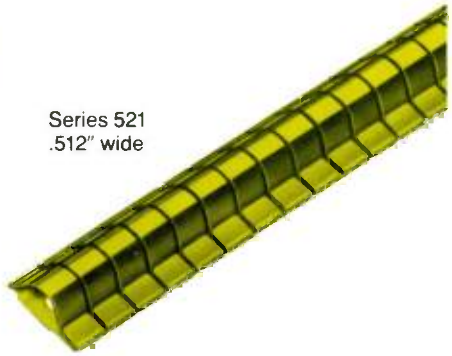

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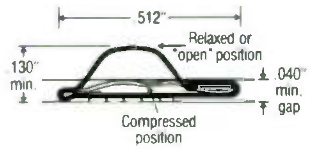

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command no. 2 to calculate the optimum terminations. The calculator informs the user that the device is not inherently stable and consequently the optimum terminations do not exist.

Since the device is not inherently stable, Stern's procedure can be used to stabilize the circuit to any desired degree. The choice of stability factor affects the gain. Typically, the stability factor should be greater than about five. Select an initial stability factor of 10 and execute routine no. 8 to generate the Y_S and Y_L given that value of k. The iterative procedure takes about 56 seconds.

How much gain will be realized? Execute command no. 3 to calculate all of the gains. The amplifier will deliver a transducer power gain of 12.77 dB which meets the specifications. The stability factor is quite large, so some of the stability could be traded for additional gain.

B. Rather than do another example of Stern's procedure for a smaller stability factor, assume that it is required that the amplifier have minimum noise figure.

The noise figure of an amplifier is a function of the source admittance. The 2N6680 transistor has a minimum noise figure of 1.60 dB with a source reflection coefficient of $0.618\angle 98$. This corresponds to a source admittance of (10.22 — j20.23) mmho. Specifying the source termination sets the available power gain. In order to realize an overall or transducer power gain equal to the available power gain, the output of the transistor must be conjugately matched.

Begin in the design by calculating the output admittance of the amplifier when the input is terminated with the source admittance for best noise figure. It is necessary to change the current value of the source admittance by executing command no. 5.

Enter the value of Y_S for best noise figure. Since the program requests Y_L also, enter some dummy value.

Next, execute command no. 4 to find out what Y_{out} will be when the input is adjusted for minimum noise figure. The calculated value of the input admittance, which is a function of the dummy value of load impedance, is ignored.

Use the change Y_S/Y_L routine again, but this time make $Y_L = Y^*_{out}$. Now, both terminations are entered in the calculator and other quantities may be determined.

First, check the stability of the circuit by executing command no. 7. The k-factor of 6.457 is adequate for good stability.

As the last step, execute command no. 3 which determines the transducer gain of the optimized amplifier. A transducer power gain of 11.85 dB is obtained with a noise figure of 1.60 dB. This completes example II.

Summary

Calculator software able to evaluate amplifier performance and to determine required source and load terminations for any condition of stability has been developed. A complete listing of the program, together with the applicable equations[1, 2] is presented in Appendix I. Appendix II contains a listing of data storage assignments. Given this information, individual subroutines can be included in user-developed programs. □

References

1. R.S. Carson, *High Frequency Amplifiers,* John Wiley, New York (1975).
2. D.R. Hertling and R.K. Feeney, "Design of RF Amplifiers Part I: Using Potentially Unstable Devices," *r.f. design,* March/April 1982, pp. 20-25. "Design of RF Amplifiers Part II: Using Inherently Stable Devices," *RF Design,* May/June 1982, pp. 30-33.


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10 dB ATTEN
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Appendix I. Program Listing.

1. Main Program

This program contains the command menu and calls subroutines required to execute a specific command.

```
01♦LBL "AMP"
02 CF 04
03 CF 03
04 ADV
05 "ENTR CMND"
06 AVIEW
07 "1-LSF"
08 AVIEW
09 "2-OPT TERM"
10 AVIEW
11 "3-GAINS"
12 AVIEW
13 "4-YIN/OUT"
14 AVIEW
15 "5-CHNG YS/L"
16 AVIEW
17 "6-CHNG YPAR"
18 AVIEW
19 "7-K-STERN"
20 AVIEW
21 "8-YS/L FOR K"
22 AVIEW
23 "9-S/Y CONV"
24 AVIEW
25♦LBL 00
26 ADV
27 FIX 0
28 "CMND?"
29 PROMPT
30 ENG 3
31 GTO IND X
32♦LBL 01
33 FC? 03
34 XEQ "PARA"
35 ADV
36 XEQ "LINV"
37 GTO 00
38♦LBL 02
39 FC? 03
40 XEQ "PARA"
41 XEQ "OPT"
42 GTO 00
43♦LBL 03
44 FC? 03
45 XEQ "PARA"
46 FC? 04
47 XEQ "YS/L"
48 XEQ "YI/O"
49 XEQ "GP"
50 XEQ "GA"
51 XEQ "GT"
52 GTO 00
53♦LBL 04
54 FC? 03
55 XEQ "PARA"
56 FC? 04
57 XEQ "YS/L"
58 XEQ "YI/O"
59 "GIN= "
60 ARCL 23
61 AVIEW
62 "BIN= "
63 ARCL 22
64 AVIEW
65 ADV
66 "GOUT= "
67 ARCL 25
68 AVIEW
69 "BOUT= "
70 ARCL 24
71 AVIEW
72 ADV
73 GTO 00
74♦LBL 05
75 XEQ "YS/L"
76 GTO 00
77♦LBL 06
78 XEQ "PARA"
79 GTO 00
80♦LBL 07
81 FC? 03
82 XEQ "PARA"
83 FC? 04
84 XEQ "YS/L"
85 XEQ "STN"
86 ADV
87 "K-STERN= "
88 ARCL 17
89 AVIEW
90 GTO 00
91♦LBL 08
92 FC? 03
93 XEQ "PARA"
94 XEQ "ITR"
95 ADV
96 XEQ "S/LPR"
97 GTO 00
98♦LBL 09
99 XEQ "SPAR"
100 XEQ "PAR1"
101 GTO 00
102 END
```

2. Linvill Stability Factor Applicable Equations:

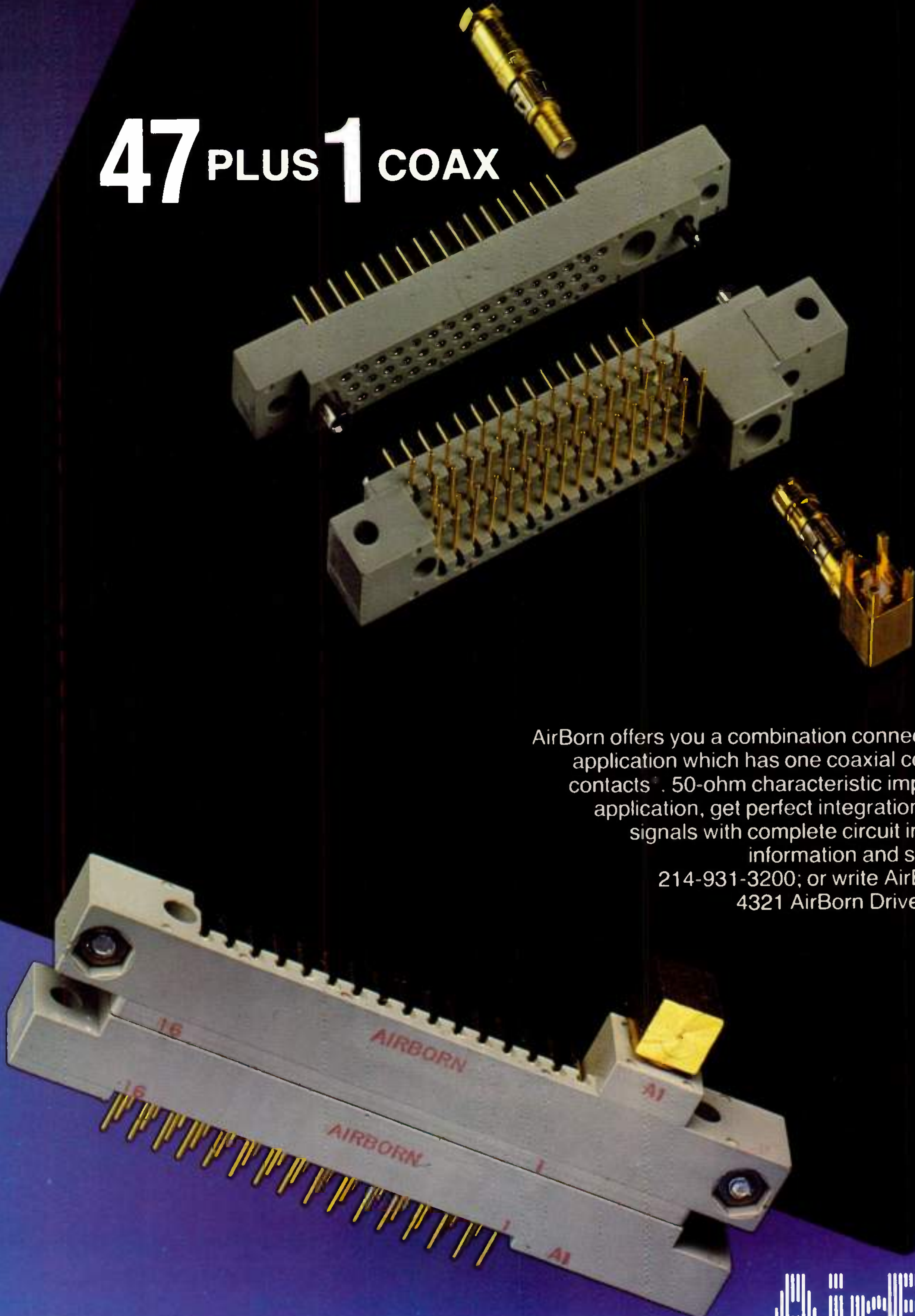
$$C = \frac{|y_r y_f|}{2g_i g_o - \mathrm{Re}(y_r y_f)}$$

Flags:

00 Set if $g_i < 0$ or $g_o < 0$ or $C < 0$ or $C \geq 1$.

```
01♦LBL "LINV"
02 CF 00
03 RCL 03
04 X<0?
05 GTO 01
06 RCL 05
07 X<0?
08 GTO 01
09 RCL 12
10 RCL 14
11 /
12 STO 16
13 X<0?
14 SF 00
15 1
16 X<>Y
17 X>Y?
18 SF 00
19 "L. S. F. = "
20 ARCL X
21 AVIEW
22 ADV
23 RTN
24♦LBL 01
25 SF 00
26 "UNSTABLE"
27 AVIEW
28 "GI/O<0"
29 AVIEW
30 END
```

47 PLUS 1 COAX



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Appendix I. Program Listing . . . Con't.

3. Optimum Terminations

Applicable Equations:

$P = \mathrm{Re}(y_r y_f)$

$Q = \mathrm{Im}(y_r y_f)$

$M = |y_r y_f|$

$$G_S)_{opt} = \frac{1}{2g_o}\sqrt{(2g_i g_o - P)^2 - M^2}$$

$$B_S)_{opt} = -b_i + \frac{Q}{2g_o}$$

$$G_L)_{opt} = \frac{1}{2g_i}\sqrt{(2g_i g_o - P)^2 - M^2}$$

$$B_L)_{opt} = -b_o + \frac{Q}{2g_i}$$

Flags:

04 set upon successful completion to indicate Y_S and Y_L generated.

```
01•LBL "OPT"
02 ADV
03 "OPTIMUM VALUES"
04 AVIEW
05 ADV
06 XEQ "LINV"
07 FS? 00
08 GTO 01
09 RCL 14
10 X↑2
11 RCL 12
12 X↑2
13 -
14 SQRT
15 STO 15
16 RCL 05
17 /
18 2
19 /
20 STO 19
21 RCL 15
22 RCL 03
23 /
24 2
25 /
26 STO 21
27 RCL 10
28 RCL 03
29 /
30 2
31 /
32 RCL 04
33 -
34 STO 20
35 RCL 16
36 RCL 05
37 /
38 2
39 /
40 RCL 02
41 -
42 STO 18
43 XEQ "S/LPR"
44 GTO 03
45•LBL "S/LPR"
46 "GS= "
47 ARCL 19
48 AVIEW
49 "BS= "
50 ARCL 18
51 AVIEW
52 ADV
53 "GL= "
54 ARCL 21
55 AVIEW
56 "BL= "
57 ARCL 20
58 AVIEW
59 ADV
60 RTN
61•LBL 03
62 RCL 08
63 RCL 09
64 R-P
65 X↑2
66 RCL 14
67 RCL 15
68 +
69 /
70 LOG
71 10
72 *
73 " dB"
74 ASTO Y
75 "GMAX= "
76 ARCL X
77 ARCL Y
78 AVIEW
79 SF 04
80 GTO 02
81•LBL 01
82 "DO NOT EXIST"
83 AVIEW
84•LBL 02
85 ADV
86 ADV
87 .END.
```

4. Power Gains

Applicable Equations:

$$G_A = \frac{|y_f|^2 G_S}{|y_i + Y_S|^2 \mathrm{Re}\{y_o - \frac{y_r y_f}{y_i + Y_S}\}}$$

$$G_P = \frac{|y_f|^2 G_L}{|y_o + Y_L|^2 \mathrm{Re}\{y_i - \frac{y_r y_f}{y_o + Y_L}\}}$$

$$G_T = \frac{4|y_f|^2 G_S G_L}{|(y_i + Y_S)(y_o + Y_L) - y_r y_f|^2}$$

Flags:

None affected.

```
01♦LBL "GA"
02 ADV
03 "AVAILABLE GAIN"
04 AVIEW
05 FS? 02
06 GTO 01
07 RCL 02
08 RCL 03
09 RCL 18
10 RCL 19
11 XEQ "Z+"
12 XEQ "MUL"
13 RCL 19
14 *
15 RCL 25
16 /
17 XEQ "DB"
18 "GA= "
19♦LBL 02
20 ARCL X
21 ARCL 28
22 AVIEW
23 RTN
24♦LBL 01
25 "DOES NOT EXIST"
26 AVIEW
27 RTN
28♦LBL "GP"
29 ADV
30 "OPERATING GAIN"
31 AVIEW
32 FS? 01
33 GTO 01
34 RCL 04
35 RCL 05
36 RCL 20
37 RCL 21
38 XEQ "Z+"
39 XEQ "MUL"
40 RCL 21
41 *
42 RCL 23
43 /
44 XEQ "DB"
45 "GP= "
46 GTO 02
47 RTN
48♦LBL "GT"
49 ADV
50 "TRANSDUCER GAIN"
51 AVIEW
52 RCL 19
53 RCL 23
54 +
55 X<0?
56 GTO 01
57 RCL 21
58 RCL 25
59 +
60 X<0?
61 GTO 01
62 RCL 02
63 RCL 03
64 RCL 18
65 RCL 19
66 XEQ "Z+"
67 STO 26
68 X<>Y
69 STO 27
70 RCL 04
71 RCL 05
72 RCL 20
73 RCL 21
74 XEQ "Z+"
75 RCL 27
76 RCL 26
77 XEQ "Z*"
78 RCL 10
79 RCL 11
80 XEQ "Z-"
81 XEQ "MUL"
82 RCL 19
83 *
84 RCL 21
85 *
86 4
87 *
88 XEQ "DB"
89 "GT= "
90 GTO 02
91♦LBL "DB"
92 LOG
93 10
94 *
95 " dB"
96 ASTO 28
97 RTN
98♦LBL "MUL"
99 XEQ "ZABS1"
100 X↑2
101 1/X
102 RCL 08
103 RCL 09
104 XEQ "ZABS1"
105 X↑2
106 *
107 END
```

5. Input and Output Admittance Applicable Equations:

$$Y_{IN} = y_i - \frac{y_r y_f}{y_o + Y_L}$$

$$Y_{OUT} = y_o - \frac{y_r y_f}{y_i + Y_S}$$

Flags:
01 set if $Re(Y_{IN}) < 0$
02 set if $Re(Y_{OUT}) < 0$

```
01♦LBL "YI/O"
02 CF 01
03 CF 02
04 1.005
05 STO 26
06 18.023
07 STO 27
08 25.019
09 STO 28
10 5.000
11 STO 29
12♦LBL 01
13 ISG 26
14 GTO 02
15 GTO 03
16♦LBL 02
17 RCL IND 26
18 ISG 26
19 RCL IND 26
20 RCL IND 27
21 ISG 27
22 RCL IND 27
23 ISG 27
24 XEQ "Z+"
25 XEQ "ZINV"
26 RCL 10
27 RCL 11
28 XEQ "Z*"
29 CHS
30 X<>Y
31 CHS
32 X<>Y
33 RCL IND 29
34 DSE 29
35 RCL IND 29
36 DSE 29
37 X<>Y
38 XEQ "Z+"
39 STO IND 28
40 DSE 28
41 X<>Y
42 STO IND 28
43 DSE 28
44 GTO 01
45♦LBL 03
46 RCL 23
47 X<0?
48 SF 01
49 RCL 25
50 X<0?
51 SF 02
52 END
```

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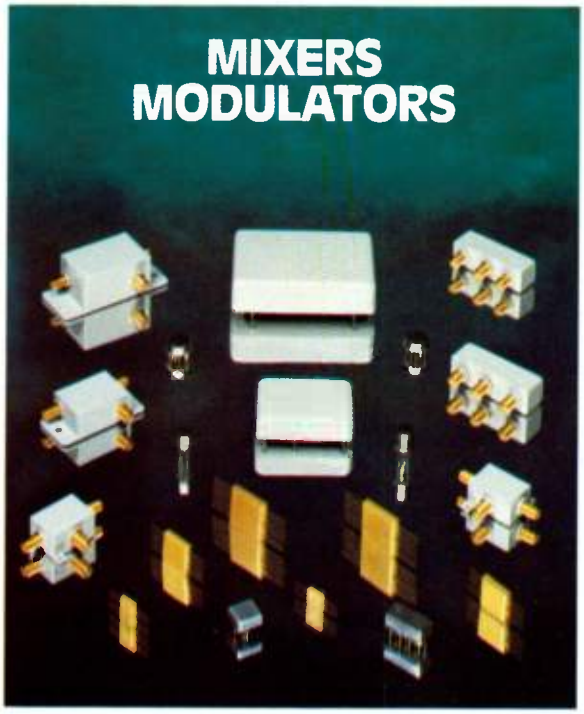

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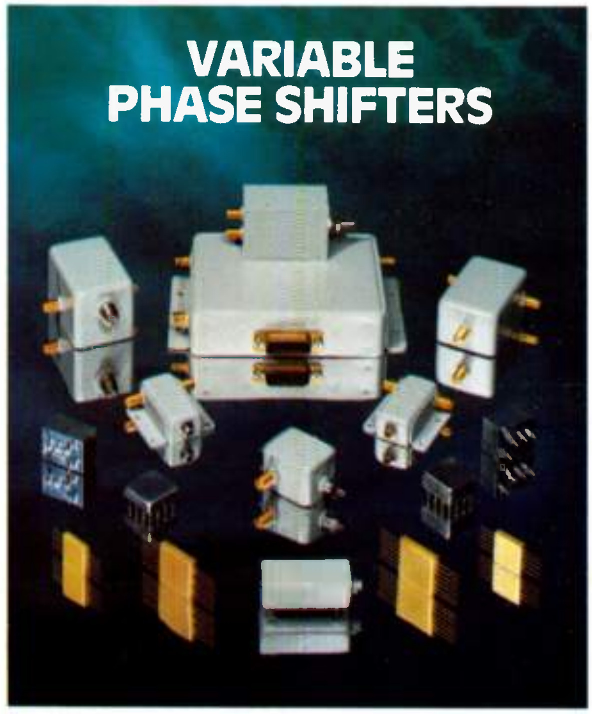

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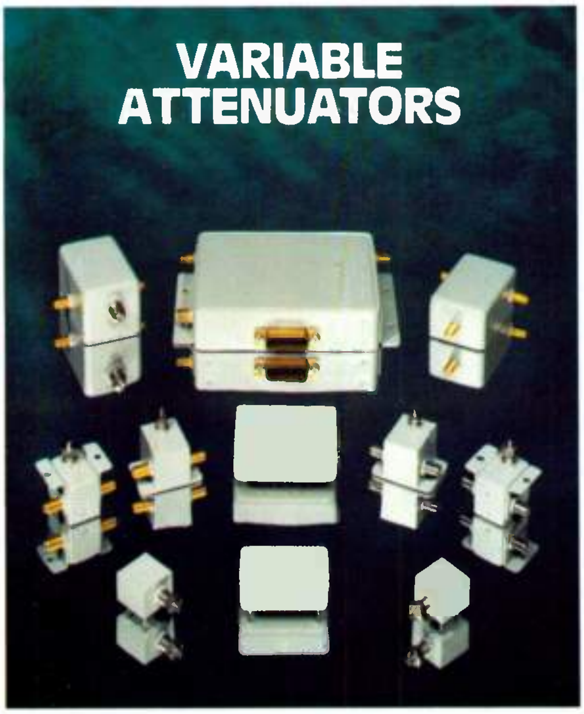

FEATURES:

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- Manually Controlled 0-20, 40 dB
- Digitally Controlled 4, 6, 8 BIT TTL
- BNC, SMA, PC Board Connections
- Panel & Non-Panel Mountable

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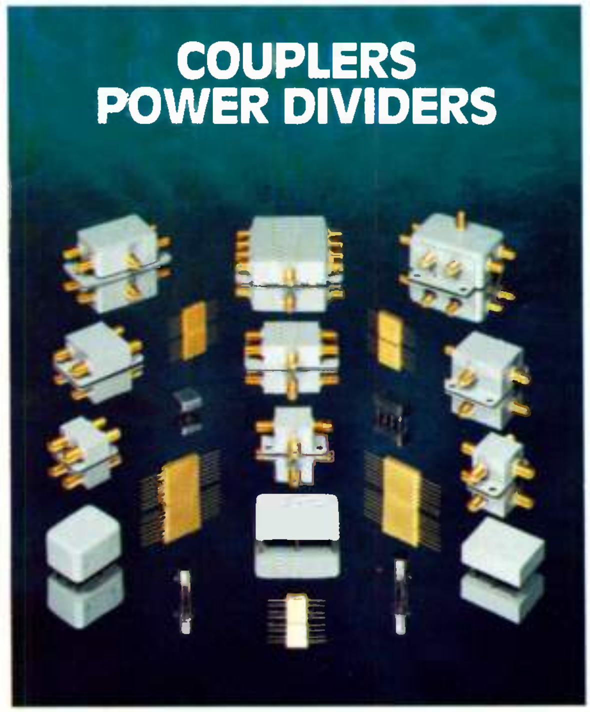

FEATURES:

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Appendix I. Program Listing . . . Con't.

6. Enter y-parameters or Y_S and Y_L

Applicable Equations:

None

Flags:

03 set when y-parameters entered

04 set when Y_S and Y_L entered

```
01•LBL "PARA"
02 ADV
03 "ENTER Y-PARAMS"
04 AVIEW
05 ADV
06 "GI=?"
07 PROMPT
08 STO 03
09 "BI=?"
10 PROMPT
11 STO 02
12 "GR=?"
13 PROMPT
14 STO 07
15 "BR=?"
16 PROMPT
17 STO 06
18 "GF=?"
19 PROMPT
20 STO 09
21 "BF=?"
22 PROMPT
23 STO 08
24 "GO=?"
25 PROMPT
26 STO 05
27 "BO=?"
28 PROMPT
29 STO 04
30•LBL "PAR1"
31 RCL 06
32 RCL 07
33 RCL 08
34 RCL 09
35 XEQ "Z*"
36 STO 11
37 X<>Y
38 STO 10
39 X<>Y
40 R-P
41 STO 12
42 RCL 03
43 RCL 05
44 *
45 2
46 *
47 STO 13
48 RCL 11
49 -
50 STO 14
51 SF 03
52 RTN
53•LBL "YS/L"
54 ADV
55 "ENTER YS AND YL"
56 AVIEW
57 "GS=?"
58 PROMPT
59 STO 19
60 "BS=?"
61 PROMPT
62 STO 18
63 "GL=?"
64 PROMPT
65 STO 21
66 "BL=?"
67 PROMPT
68 STO 20
69 SF 04
70 END
```

7. Calculate Stern Stability Factor or Specify Terminations for Given k

Applicable Equations:

$P = R_e(y_r y_f)$

$Q = I_m(y_r y_f)$

$M = |y_r y_f|$

To calculate k:

$$k = \frac{2 g_i g_o \left(1 + \frac{G_S}{g_i}\right)\left(1 + \frac{G_L}{g_o}\right)}{P + M}$$

To specify terminations:

if $g_i > 0, g_o > 0$

then,

$$\frac{G_S}{|g_i|} = \frac{G_L}{|g_o|} = \sqrt{\frac{k(P + M)}{2|g_i g_o|}} - 1$$

if $g_i < 0, g_o > 0$

or $g_i > 0, g_o < 0$

then,

$$\frac{G_S}{|g_i|} = \frac{G_L}{|g_o|} = \sqrt{\frac{k(P + M)}{2|g_i g_o|} - 1}$$

if $g_i < 0, g_o < 0$

then,

$$\frac{G_S}{|g_i|} = \frac{G_L}{|g_o|} = \sqrt{\frac{k(P + M)}{2|g_i g_o|}} + 1$$

Flags:
When calculating k,
07 set if either or both
$(g_i + G_S) < 0$
$(g_o + G_L) < 0$
When calculating terminations, 05 and 06 are set when the convergence is complete.

```
01♦LBL "SPAR"
02 "MS11=?"
03 PROMPT
04 STO 31
05 "AS11=?"
06 PROMPT
07 STO 30
08 "MS12=?"
09 PROMPT
10 STO 35
11 "AS12=?"
12 PROMPT
13 STO 34
14 "MS21=?"
15 PROMPT
16 STO 37
17 "AS21=?"
18 PROMPT
19 STO 36
20 "MS22=?"
21 PROMPT
22 STO 33
23 "AS22=?"
24 PROMPT
25 STO 32
26 29.037
27 STO 26
28♦LBL 01
29 ISG 26
30 GTO 02
31 GTO 03
32♦LBL 02
33 RCL IND 26
34 ISG 26
35 RCL IND 26
36 P-R
37 STO IND 26
38 1
39 ST- 26
40 RDN
41 X<>Y
42 STO IND 26
43 ISG 26
44 GTO 01
45♦LBL 03
46 RCL 34
47 RCL 35
48 RCL 36
49 RCL 37
50 XEQ "Z*"
51 STO 28
52 X<>Y
53 STO 27
54 1
55 RCL 31
56 +
57 STO 29
58 RCL 30
59 X<>Y
60 1
61 RCL 33
62 +
63 STO 38
64 RCL 32
65 X<>Y
66 XEQ "Z*"
67 RCL 27
68 RCL 28
69 XEQ "Z-"
70 STO 42
71 X<>Y
72 STO 41
73 2
74 RCL 29
75 -
76 STO 39
77 2
78 RCL 38
79 -
80 STO 40
81 RCL 30
82 CHS
83 RCL 39
84 RCL 32
85 RCL 38
86 XEQ 05
87 STO 03
88 X<>Y
89 STO 02
90 RCL 30
91 RCL 29
92 RCL 32
93 CHS
94 RCL 40
95 XEQ 05
96 STO 05
97 X<>Y
98 STO 04
99 -2
100 RCL 34
101 *
102 -2
103 RCL 35
104 *
105 RCL 41
106 RCL 42
107 XEQ "Z/"
108 STO 07
109 X<>Y
110 STO 06
111 -2
112 RCL 36
113 *
114 -2
115 RCL 37
116 *
117 RCL 41
118 RCL 42
119 XEQ "Z/"
120 STO 09
121 X<>Y
122 STO 08
123 2.009
124 STO 26
125♦LBL 04
126 .02
127 RCL IND
128 *
129 STO IND
130 CHS
131 ISG 26
132 GTO 04
133 RTN
134♦LBL 05
135 XEQ "Z*"
136 RCL 27
137 RCL 28
138 XEQ "Z+"
139 RCL 41
140 RCL 42
141 XEQ "Z/"
142 END
```

8. S-Parameter to y-Parameter Conversion
Applicable Equations:

$$y_i = \frac{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$$

$$y_o = \frac{(1 + S_{11})(1 - S_{22}) + S_{12}S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$$

$$y_r = \frac{-2S_{12}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$$

$$y_f = \frac{-2S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$$

$$y = y' \left(\frac{1}{z_o}\right),$$

where z_o is the S-parameter reference impedance, $1/z_o$ is set to 0.02 in the program (step 126).
Flags:
03 set when y-parameters are calculated.

```
01♦LBL "STN"
02 CF 07
03 RCL 03
04 RCL 19
05 +
06 X<0?
07 SF 07
08 RCL 05
09 RCL 21
10 +
11 X<0?
12 SF 07
13 *
14 2
15 *
16 RCL 11
17 RCL 12
18 +
19 /
20 STO 17
21 FC? 07
22 RTN
23 ABS
24 CHS
25 STO 17
26 RTN
27♦LBL "ITR"
28 "K=?"
29 PROMPT
30 STO 17
31 RCL 11
32 RCL 12
33 +
34 *
35 2
36 /
37 RCL 03
38 RCL 05
39 *
40 ABS
41 /
42 STO 27
43 RCL 03
44 RCL 05
45 *
46 X<0?
47 GTO 02
48 RCL 03
49 X<0?
50 GTO 03
51 RCL 27
52 SQRT
53 1
54 -
55 GTO 05
56♦LBL 02
57 RCL 27
58 1
59 +
60 SQRT
61 GTO 05
62♦LBL 03
63 RCL 27
64 SQRT
65 1
66 +
67♦LBL 05
68 STO 26
69 RCL 03
70 ABS
71 *
72 STO 19
73 RCL 26
74 RCL 05
75 ABS
76 *
77 STO 21
78 CF 05
79 CF 06
80 RCL 02
81 STO 22
82 RCL 04
83 STO 24
84♦LBL 01
85 RCL 22
86 CHS
87 STO 18
88 RCL 24
89 CHS
90 STO 20
91 XEQ "YI/O"
92 RCL 22
93 RCL 18
94 +
95 RCL 22
96 /
97 "ABS"
98 1 E-4
99 X>Y?
100 SF 05
101 RCL 24
102 RCL 20
103 +
104 RCL 24
105 /
106 ABS
107 1 E-4
108 X>Y?
109 SF 06
110 FC? 05
111 GTO 01
112 FC? 06
113 GTO 01
114 SF 04
115 BEEP
116 END
```

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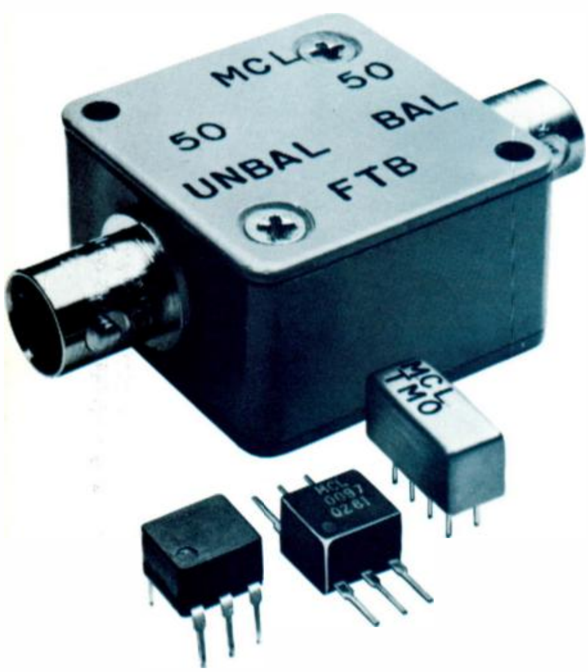

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Model No.	T1-1 TMO1-1	T1-1H	T1.5-1 TMO1.5-1	T2.5-6 TMO2.5-6	T4-6 TMO4-6	T9-1 TMO9-1	T9-1H	T16-1 TMO16-1	T16-1H
Imped. Ratio	1	1	1.5	2.5	4	9	9	16	16
Freq. (MHz)	.15-400	8-300	.1-300	.01-100	.02-200	.15-200	2-90	.3-120	7-85
T Model (10-49)	$2.95	$4.95	$3.95	$3.95	$3.95	$3.45	$5.45	$3.95	$5.95
TMO model (10-49)	$4.95		$6.75	$6.45	$6.45	$6.45		$6.45	

CENTER-TAPPED DC ISOLATED PRIMARY & SECONDARY

Model No.	T1-1T TMO1-1T	T2-1T TMO2-1T	T2.5-6T TMO2.5-6T	T3-1T TMO3-1T	T4-1 TMO4-1	T4-1H	T5-1T TMO5-1T	T13-1T TMO13-1T
Imped. Ratio	1	2	2.5	3	4	4	5	13
Freq. (MHz)	.05-200	.07-200	.01-100	.05-250	.2-350	8-350	.3-300	.3-120
T Model (10-49)	$3.95	$4.25	$4.25	$3.95	$2.95	$4.95	$4.25	$4.25
TMO model (10-49)	$6.45	$6.75	$6.75	$6.45	$4.95		$6.75	$6.75

UNBALANCED PRIMARY & SECONDARY

Model No.	T2-1 TMO2-1	T3-1 TMO3-1	T4-2 TMO4-2	T8-1 TMO8-1	T14-1 TMO14-1
Imped. Ratio	2	3	4	8	14
Freq. (MHz)	.025-600	5-800	.2-600	.15-250	.2-150
T model (10-49)	$3.45	$4.25	$3.45	$3.45	$4.25
TMO Model (10-49)	$5.95	$6.95	$5.95	$5.95	$6.75

FT FTB

Model No.	FT1.5-1	FTB1-1	FTB1-6	FTB1-1-75
Imped. Ratio	1.5	1	1	1
Freq. (MHz)	.1-400	.2-500	.01-200	.5-500
(1-4)	$29.95	$29.95	$29.95	$29.95

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Appendix I. Program Listing . . . Con't.

9. Complex Arithmetic

Routines do complex addition, subtraction, multiplication, division and absolute value. Most are similar to those in the HP Math Pac I module.

Applicable Equations:
Standard complex arithmetic.

Flags:
None

```
01♦LBL "Z+"
02 X<>Y
03♦LBL 05
04 RDN
05 +
06 RDN
07 +
08 R↑
09 RTN
10♦LBL "Z-"
11 CHS
12 X<>Y
13 CHS
14 GTO 05
15♦LBL "ZINV"
16 XEQ 02
17 RTN
18♦LBL 02
19 R-P
20 1/X
21 X<>Y
22 CHS
23 X<>Y
24 P-R
25 RTN
26♦LBL "Z*"
27 R-P
28 RDN
29 RDN
30 R-P
31 RDN
32 RDN
33 X<>Y
34 RDN
35 *
36 RDN
37 +
38 R↑
39 P-R
40 RTN
41♦LBL "Z/"
42 XEQ 02
43 GTO "Z*"
44♦LBL "ZABS1"
45 R-P
46 RDN
47 RDN
48 RCL Z
49 END
```

Appendix II. Storage Assignments.

Register	Quantity	Register	Quantity	Reg.	Qty.
00	Not used	18	B_S } Y_S	36	Im(S21)
01	Not used	19	G_S	37	Re(S21)
02	b_i } y_i	20	B_L } Y_L	38	scratch
03	g_i	21	G_L	39	scratch
04	b_o } y_o	22	B_{in} } Y_{IN}	40	scratch
05	g_o	23	G_{in}	41	scratch
06	b_r } y_r	24	B_{out} } Y_{OUT}	42	scratch
07	g_r	25	G_{out}		
08	b_f } y_f	26	scratch	registers 30-42 used only by S-parameter routine	
09	g_f	27	scratch		
10	$Q = Im(y_r y_f)$	28	scratch		
11	$P = Re(y_r y_f)$	29	scratch		
12	$M = \lvert y_r y_f \rvert$	30	Im(S11)		
13	$2g_i g_o$	31	Re(S11)		
14	$2g_i h_o - P$	32	Im(S22)		
15	$(2g_i g_o - P)^2 - M^2$	33	Re(S22)		
16	C(Linvill factor)	34	Im(S12)		
		35	Re(S12)		
17	k (Stern factor)				

Lumped-Constant Line Stretcher For Testing Power Amplifier Stability

To determine the stability of an RF power amplifier, oscillator, or transmitter under harsh conditions, a variable line and fixed attenuator are used to simulate a given VSWR at all phase angles. This article describes a lumped-constant line stretcher using either variable capacitors or inductors.

By E.A. Franke and A.E. Noorani
General Electric Company
Lynchburg, VA 24502

Transmitters employing electron tubes were typically narrowband tuned devices. With the advent of the transistor, the RF power amplifier is capable of wideband operation and must therefore be evaluated for stability to avoid interference with other signals. The spectral purity must be maintained in spite of variations in temperature, supply voltage, or load impedance. For applications such as marine, automotive, or portable operation, the antenna may present a severe mismatch to the power amplifier due to the proximity of foreign objects, environmental corrosion, or a pinched coaxial cable. A most stringent test of stability occurs when the power amplifier operates into a high-Q load such as a narrowband duplexer or cavity filter. Commercial mobile transmitters are typically rated to maintain stability into a 3:1 VSWR even under fluctuations of ± 20% in supply voltage.

The power amplifier may be tested into a worse case mismatch at all phase angles. This is easily simulated by connecting the amplifier output to an attenuator followed by an adjustable air line terminated in either an open or a short, (Figure 1). Stability is monitored on a spectrum analyzer connected to a directional coupler placed between the power amplifier and the attenuator. Movement of the air line through one-half wavelength will traverse a complete circle around the Smith Chart. The radius of the circle will depend on the size of attenuator used. The mismatch presented by the artificial load is best described using return loss. The value of return loss is simply twice the attenuation of the fixed pad, assuming negligible loss in the adjustable air line. The mismatch in terms of standing wave ratio is given as

$$\text{VSWR} = \coth\left[\frac{\text{attenuator (dB)}}{8.686}\right] \quad (1)$$

The radius of the mismatch circle on the normalized Smith Chart intersects the right half of the abscissa at the value of the VSWR.

Recently Roderick Blocksome[1] reported on a binary stepped transmission line formed by numerous lengths of coaxial cable and relays. Below 200 MHz the length of a half-wave adjustable air line (>75 cm) becomes so cumbersome that a continuously-variable, lumped-constant version is needed. This article suggests a stretcher using either variable capacitors or inductors instead of transmission lines.

Lumped-Constant Line Stretcher

Consider the network shown in Figure 2. The input impedance is simply the parallel combination of two series circuits L1, C1 and L2, C2.

$$Zin = j\frac{\left(\omega L_1 - \frac{1}{\omega C_1}\right)\left(\omega L_2 - \frac{1}{\omega C_2}\right)}{\omega(L_1 + L_2) - \frac{1}{\omega}\left(\frac{1}{C_1} + \frac{1}{C_2}\right)} \quad (2)$$

where $\omega = 2\pi f$.

PAD	RETURN LOSS	VSWR
1 dB	2 dB	8.72:1
2 dB	4 dB	4.42:1
3 dB	6 dB	3.00:1
4 dB	8 dB	2.32:1

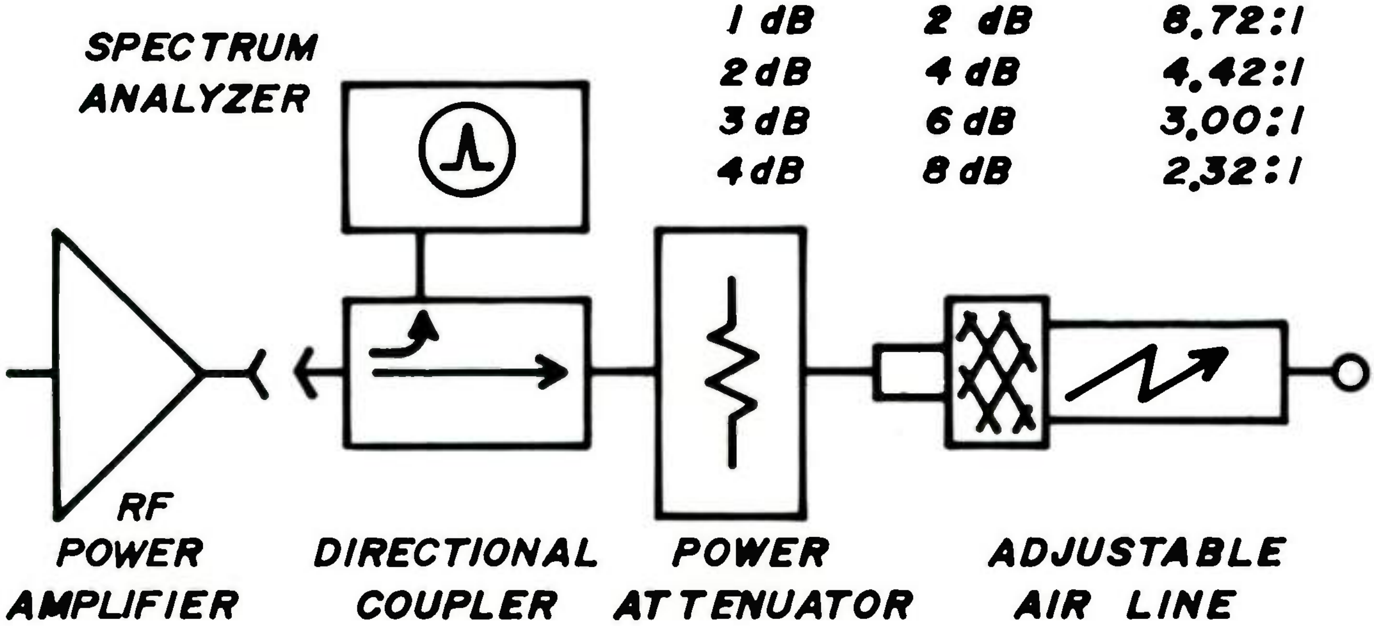

Figure 1. The spectral stability of a power amplifier may be examined for a mismatch at all phase angles.

Examination of the above equation shows two zeros in the numerator corresponding to two series resonances. A parallel resonance occurs in the denominator when

$$\omega(L_1 + L_2) = \frac{2}{\omega C} , \qquad (3)$$

where $C_1 = C_2 = C$ and assuming a ganged dual section or butterfly variable capacitor. The above equation may be rewritten as

$$XC = \frac{XL_1 + XL_2}{2}$$

where $XL_1 = \omega L_1$, $XL_2 = \omega L_2$ and $X_C = \frac{1}{\omega C}$.

At any given frequency, the locus of impedance points plotted on a Smith Chart as the capacitor is varied will be a circle similar to that achieved using a variable length transmission line. If an attenuator is not used, the points will be entirely imaginary (R = 0). If an attenuator is used, the value of the real and imaginary parts of the input impedance may simply be read from the chart. When properly designed, as the capacitance is increased, the locus travels from a short circuit (when $XL_1 = X_C$) (Figure 3) to an inductive

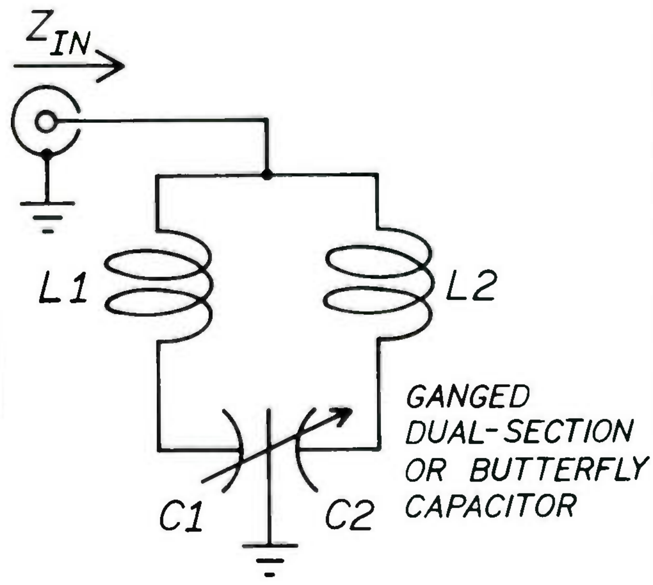

Figure 2. The lumped-constant line stretcher is formed by two series circuits which combine for a parallel combination.

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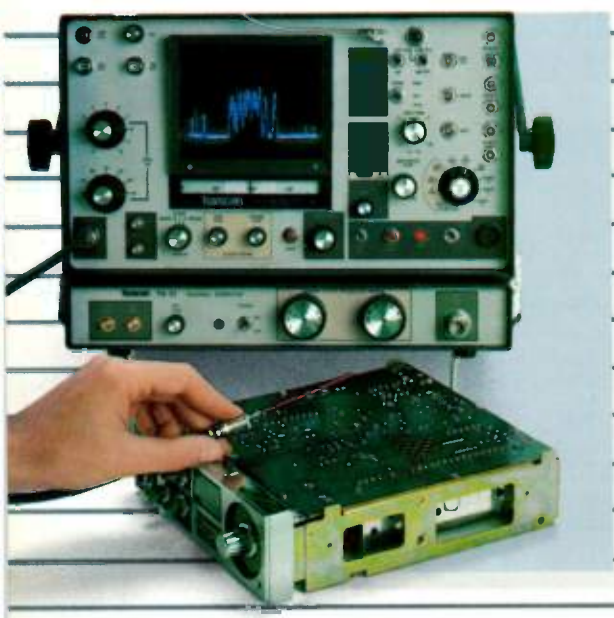

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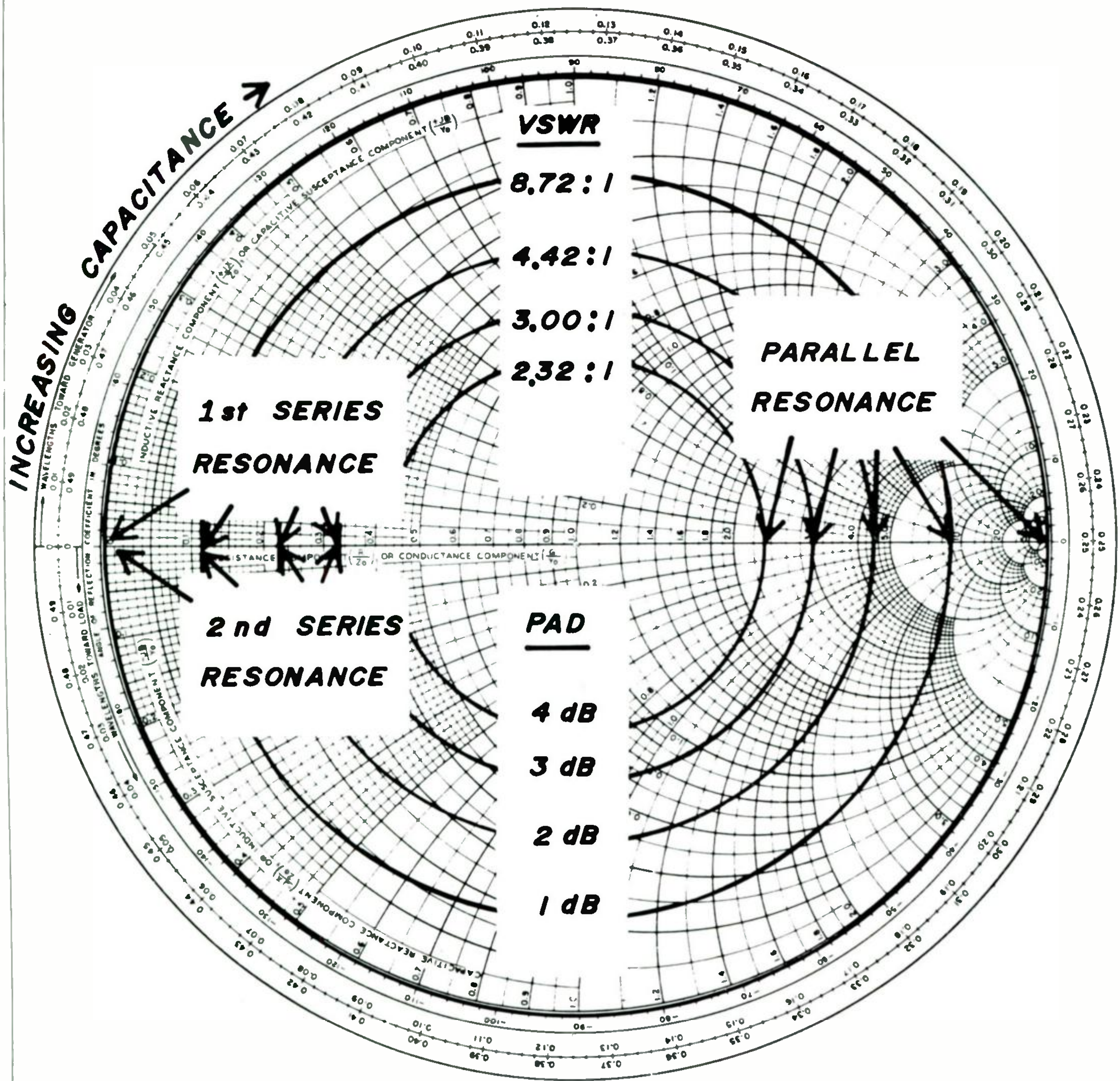

Figure 3. As the variable capacitor is meshed, the locii of input impedance points forms a clockwise circle of constant VSWR covering all phase angles.

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reactance, then through parallel resonance to a capacitive reactance. Finally the locus passes through the second series resonance (when $XL_2 = X_C$) near maximum capacitance.

By properly dictating the end points of the locii at the series resonances, the parallel resonance will occur somewhere between them.

For design purposes, it is necessary to establish the frequency bandwidth over which the lumped-constant variable line has to work. Next the ganged dual-section or butterfly capacitor selected must have a high breakdown voltage and a very high Q. Any dissipative loss in the coil or capacitor will be evidenced by a non-circular response on the Smith Chart. Before any calculations for the inductors are made, the minimum and maximum capacitance should be measured. It is also necessary that the capacitor does not have a self resonance anywhere in the frequency band of interest.

Inductor L_1 is calculated using the maximum capacitance at the lowest operating frequency for series resonance.

$$L_1 = \frac{1}{4\pi^2(F_{low})^2 C_{max}} \qquad (4)$$

The second inductor L_2 is calculated using the minimum variable capacitance at the highest operating frequency.

$$L_2 = \frac{1}{4\pi^2(F_{high})^2 C_{min}} \qquad (5)$$

The ratio of inductance L1/L2 must be less than unity

$$\frac{L1}{L2} = \left(\frac{F_{high}}{F_{low}}\right)^2 \left(\frac{C_{min}}{C_{max}}\right), \qquad (6)$$

which reduces to

$$\left(\frac{F_{high}}{F_{low}}\right)^2 < \left(\frac{C_{max}}{C_{min}}\right). \qquad (7)$$

If the inductors are equal, then parallel resonance will occur at the same time as series resonance both at the highest frequency (C_{min}) and at the lowest frequency (C_{max}).

If the ratio of L1/L2 is greater than unity, parallel resonance will not occur within the frequency range F_{low} to F_{high}.

From equation 7, the operating range is limited by the square root of the capacitance ratio. It should be remembered that the input impedance will vary faster when a wider frequency range is chosen. Thus to insure a more gradual change in impedance as the capacitor is rotated, the frequency range is usually limited only to the band of interest.

VHF Example

As an example calculation, a variable line was designed to operate over the two-way radio band, 150 to 175 MHz. A dual variable capacitor was chosen with a minimum capacitance of 7.5 pF and a maximum of 50 pF. We shall limit the range of capacitance to C_{min} = 10 pF and C_{max} = 45 pF to account for parasitic capacitance and to operate slightly away from the capacitor stops.

$$L_1 = \frac{1}{4\pi^2 (F_{low})^2 C_{max}} = \frac{1}{4\pi^2 (150\text{MHz})^2 (45\text{pF})} = 25 \text{ nH}$$

$$L_2 = \frac{1}{4\pi^2 (F_{high})^2 C_{min}} = \frac{1}{4\pi^2 (175\text{ MHz})^2 (10\text{pF})} = 83 \text{ nH}$$

Very low values of inductance may be simply short lengths of wire or flat copper strips. The exact value is determined by adjusting for series resonance at each end of the capacitor travel.

The input impedance may be quickly examined at band edges. At 150 MHz the first series resonance occurs at C_{max} = 45pF by the above design. At the second series resonance, the variable capacitor resonates with the other inductor L2.

$$X_C = X_{L_2} = 2\pi f L_2 = 2\pi(150\text{ MHz})(83\text{ nH}) = 78 \text{ ohms}$$

$$C = \frac{1}{2\pi f X_c} = \frac{1}{2\pi(150\text{MHz}(78\text{ ohm})} = 13.6\text{ pF}$$

A parallel resonance will occur between the two series resonances when all of the components are taken into account.

$$XC = \frac{XL_1 + XL_2}{2} = \frac{2\pi f(L_1 + L_2)}{2}$$

$$= \pi(150\text{MHz})(25\text{nH} + 83\text{nH}) = 50.9 \text{ ohm}$$

$$C = \frac{1}{2\pi f XC} = \frac{1}{2\pi(150\text{MHz})(78\text{ ohm})} = 20.8\text{ pF}$$

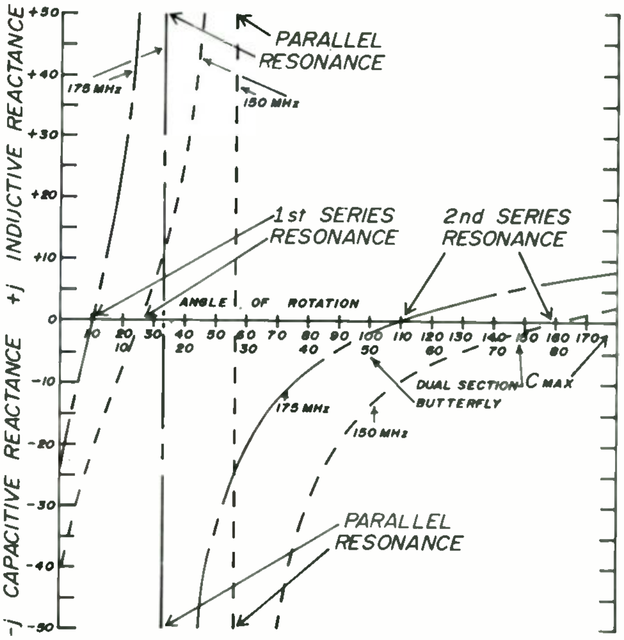

Figure 4.

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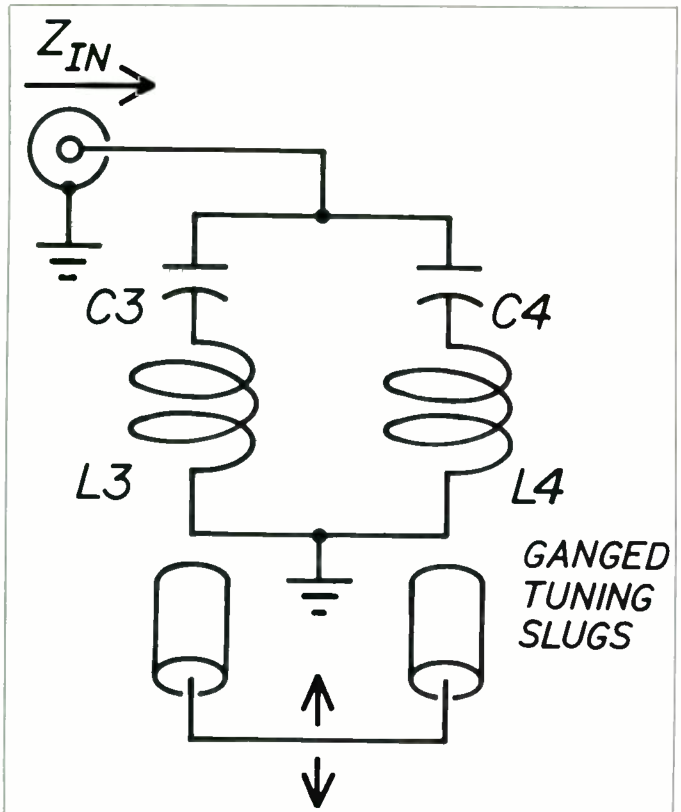

Figure 5. The lumped-constant adjustable line simulator may also be formed using a permeability tuner.

At the high end of the band (175 MHz) the following conditions are present:

First Series Resonance C = 33 pF
Parallel Resonance C = 15.3 pF
Second Resonance C = 10 pF

The input impedance is plotted as a function of capacitor rotation for the lowest and highest operating frequency in Figure 4. Rotation is shown both for a semi-circular (180°) and a butterfly (90°) capacitor. A gear reduction is typically employed with a calibrated dial. The equation for the capacitor is assumed to be linear

$$C = [(\theta/180^\circ)(C_{max} - C_{min})] + C_{min}$$

where

θ = angular rotation in degrees from fully open
C_{max} = 50 pF
C_{min} = 7.5 pF

The allowable RF power input to this type of device is dependent on the breakdown voltage of the capacitor. For the above design a 500V breakdown capacitor was used and 150 Watts of RF power was applied without any adverse effects.

Variable Inductance Line Stretcher

The variable may be switched from the capacitor to the inductor (Figure 5). A slug formed from ferrite, brass, copper, or aluminum may be mechanically positioned within two identical inductors to vary the inductance. The fixed capacitors, C3 and C4, are computed in a similar manner as above. The conductive metallic core acts as a single shorted turn to decrease the inductance with increased penetration. The powdered iron or ferrite core has the opposite effect and increases the inductance as it penetrates the coil. The Q of the coil is not appreciably lowered when using the brass slug provided the slug has a clean surface or is silver plated. The use of the powdered iron or ferrite core will raise the Q of the coil provided the material has been selected for the frequency range in use. Ferrite and powdered iron are typically usable up to 200 MHz.

Variable Inductor Example

When a slug having a permeability of μ is inserted into an air-core inductance L_{air}, the final inductance becomes $\mu L_{air} = L_{core}$. Thus L_{core} and L_{air} are the maximum and minimum values of inductance.

If the minimum and maximum of the matched coils are chosen for example to be $L_{max} = L_{core} = 83$ nH and $L_{min} = L_{air} = 25$ nH, the values of capacitance may be then calculated.

$$\mu = \frac{L_{core}}{L_{air}} = \frac{83\text{ nH}}{25\text{ nH}} = 3.3\text{, ferrite}$$

$$C_3 = \frac{1}{4\pi^2 (F_{low})^2 L_{max}}$$

$$C_4 = \frac{1}{4\pi^2 (F_{high})^2 L_{min}}$$

For the frequency range of 150 to 174 MHz, $C_3 = 13.56$ pF and $C_4 = 33.08$ pF. Parallel resonance will occur when

$$2\omega L = \frac{1}{\omega}\left(\frac{1}{C_3} + \frac{1}{C_4}\right)$$

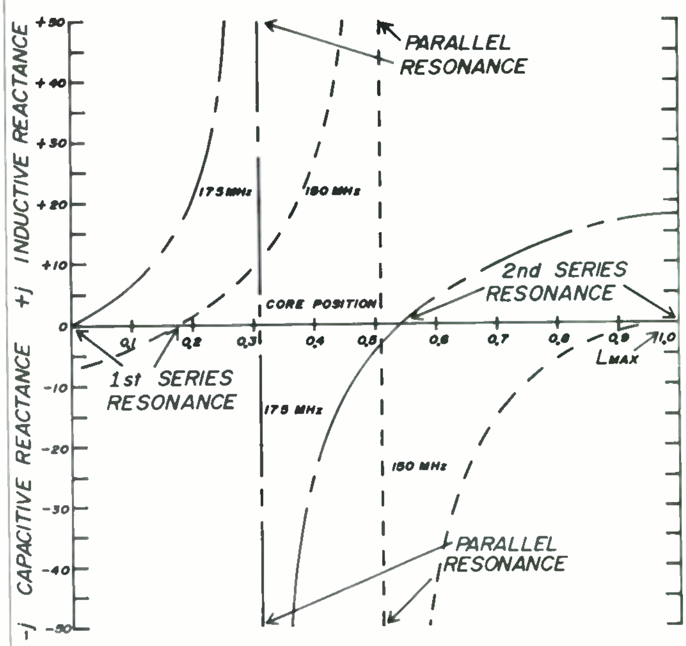

Figure 6. The input for the variable inductor is similar to the previous case as the ferrite slug is inserted into the coil.

At the low frequency of 150 MHz the value of each inductor will be

$$L = \frac{C_3 + C_4}{2\omega^2 C_3 C_4} = 58.5 \text{ nH}$$

for parallel resonance. The value of each inductor will be 43 nH for parallel resonance at 175 MHz. This is plotted in Figure 6. The abscissa is plotted as linear travel for the position of a ferrite slug into a cylindrical, linearly-wound coil whose dimensions equal those of the core being used. It represents the fraction of the core inside the coil.

Conclusion

To determine the stability of an RF power amplifier, oscillator, or transmitter under harsh conditions, a variable line and fixed attenuator are used to simulate a given VSWR at all phase angles. Below 200 MHz a half-wave adjustable line is cumbersome. This article has described a lumped-constant line stretcher using either variable capacitors or inductors.

The use of moveable slugs eliminates mechanical wear at the electrical grounding point present on the capacitor rotor. For automated testing of production transistors or amplifiers, a frictionless device is necessary.

Today there are few manufacturers still fabricating dual section or butterfly variable capacitors because their use was primarily in electron tube amplifiers.

Reference

1. R.K. Blocksome, "A Binary Stepped Transmission Line," *r.f. design*, July/August 1982, pp. 22-29. □

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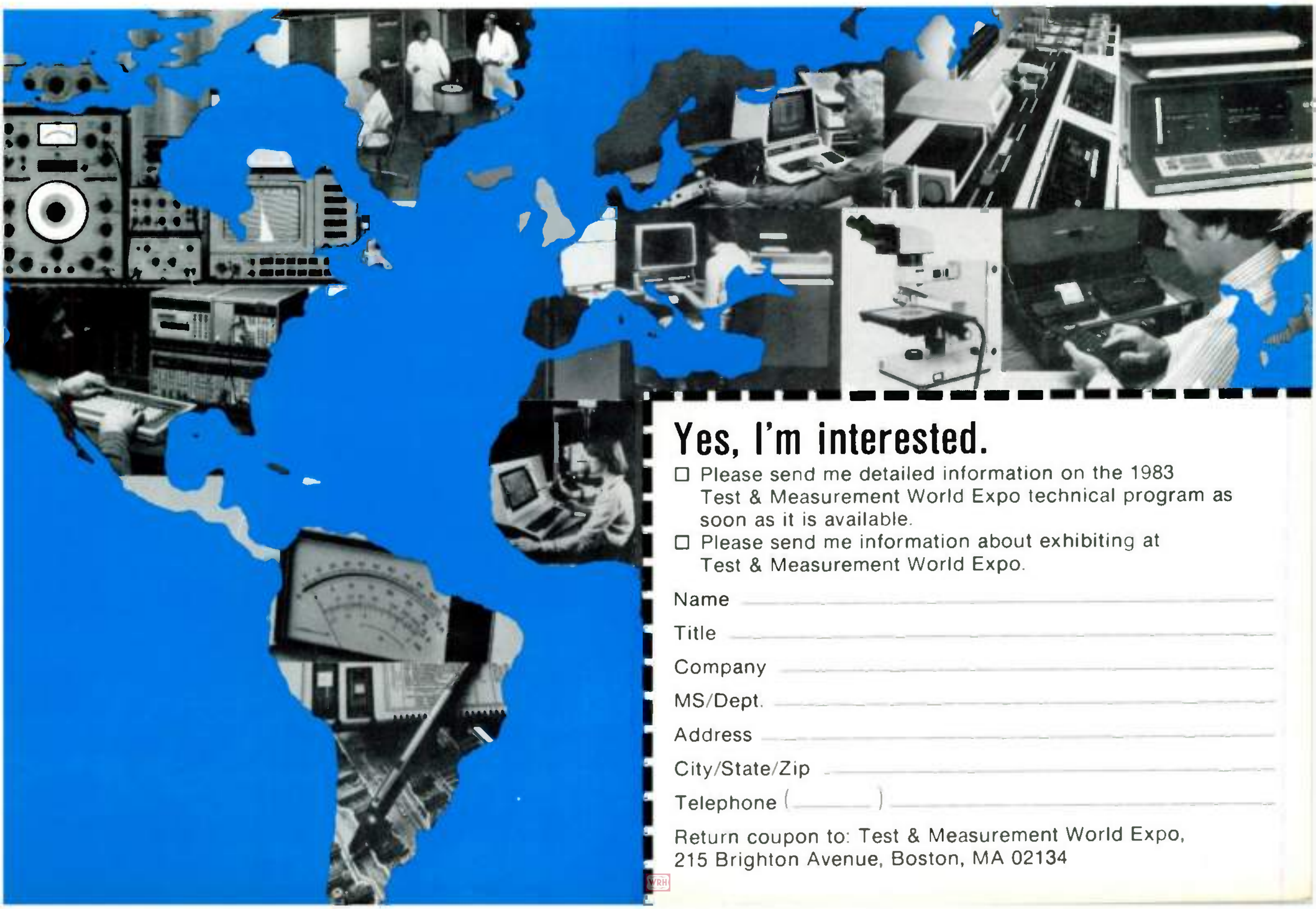

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An Easy Method for Measuring Unloaded Q (Q_u) for an Inductor or Filter Tank

A measurement technique to measure inductance by using a narrowband single pole bandpass filter whose Q approximately equals the unloaded Q of the inductor or tank being measured.

By Marvin Kefer
Hazeltine Corp.
Greenlawn, N.Y.

This article presents an easy, accurate method for selecting the best (highest Qu) inductor or tank before you build a filter. Measurements are done at the operating frequency and include self resonant effects. The method of measurement uses readily available laboratory equipment, and is usable at any frequency.

The described technique uses a measurement circuit which is a narrowband, single-pole, bandpass filter whose inductance (or tank) will be measured. Conceptually, the filter Q is dependent on (and limited by) the unloaded Q of the tank or inductor. By measuring the filter's Q, the tank or inductor is defined. The method is illustrated by an example which calculates the inductance and then resolves the inductor, or tank Q, from the filter's Q.

Construction of the Measurement Circuit

This measurement method requires that the resonated inductor, or tank, be lightly coupled to the source and load as shown in Figure 1. At low frequencies the coupling can be capacitive or inductive; at higher frequencies magnetic or electric probes are suggested.

A 285 MHz bandpass filter design requires a 75nH inductor. The measurement of inductance and Qu uses the test circuit shown in Figure 2. Light coupling and resonant capacitors are implemented by using accurately measured high Q series chip capacitors. It is tacitly presumed that Qu (coupling

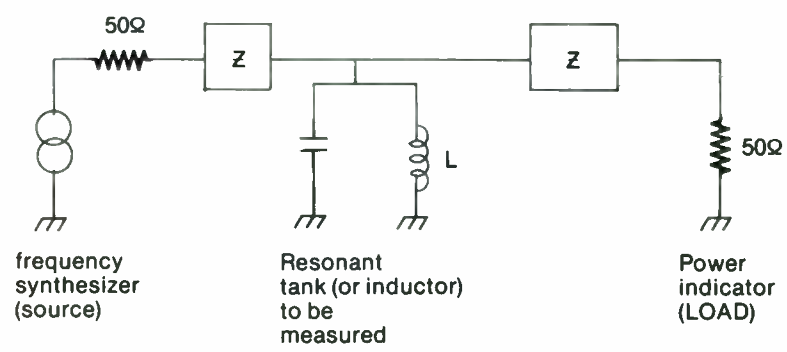


Figure 1. The source and load are lightly coupled (electrically or magnetically), by z, to the Resonant Tank.

capacitor) > Qu (resonant capacitor) > Qu (inductor) and therefore the effects of these very high capacitor Qu's will be ignored: The parallel resonance equivalent circuit which is used to illustrate the calculations is shown in Figure 2. The series to parallel transformation (derived Appendix I) is used to convert the test circuit to the parallel resonant circuit in Appendix II.

Measurement of the 3dB bandpass frequencies is used to find Q and L as shown below (see Figure 3).

1. $f_o = \sqrt{f_1 f_2} \approx (f_1 + f_2)/2$ (narrowband approximation)
   where f_1 = lower 3dB frequency (283.5 MHz)
   f_2 = higher 3dB frequency (286.4 MHz)
   = 284.95 MHz
2. $Q = f_0/BW(3dB)$
   where $BW(3dB) = f_2 - f_1 =$
   = 2.9 MHz
   = 98.26
3. $L = 1/(2\pi f_0)^2 C$
   where C = (3.96 + 0.1 + 0.1) pF (from Figure 2)
   = 75 nH

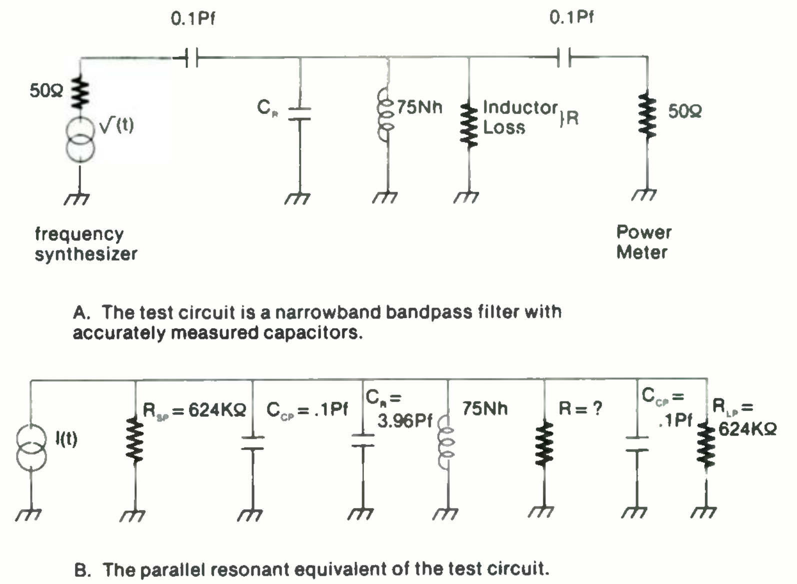
It should be noted that synthesizer frequency is usually very accurately displayed and measured. Overall measurement accuracy is dependent on the accuracy of C (capacitance) measurement and of the 3dB insertion loss measurement.

Figure 3 shows that the 3dB frequencies have well defined insertion loss. Therefore, they can be used to calculate f_0, instead of directly measuring f_0, whose insertion loss is not well defined for small variations of f_0 (Passband frequencies).

The Q calculated is equal to Qu of the inductor or tank when the insertion loss of the test circuit is high as derived in Appendix III (shown in Figure 4). To improve the accuracy for Qu, we measure the insertion loss at f_0 (Figure 3). The graph of the relationship between Qu and Q is shown in Figure 4. For example, the bandpass insertion loss of f_0 in Figure 3 (20 dB) when applied to Figure 4 indicates Qu/Q = 1.1 or

4. Qu = (1.1)(98.26) = 108.

The 20 dB insertion loss is an actual measurement and indicates circuit perturbations due to circuit stray coupling capacitance. Stray capacitance affects the calculated inductance value, but not the value of Qu because the relationship between insertion loss (IL), filter Q and inductance (or tank) Qu (given by Equation



A. The test circuit is a narrowband bandpass filter with accurately measured capacitors.

B. The parallel resonant equivalent of the test circuit.

Figure 2. The parallel equivalent circuit for the test circuit is used as a basis for calculations.

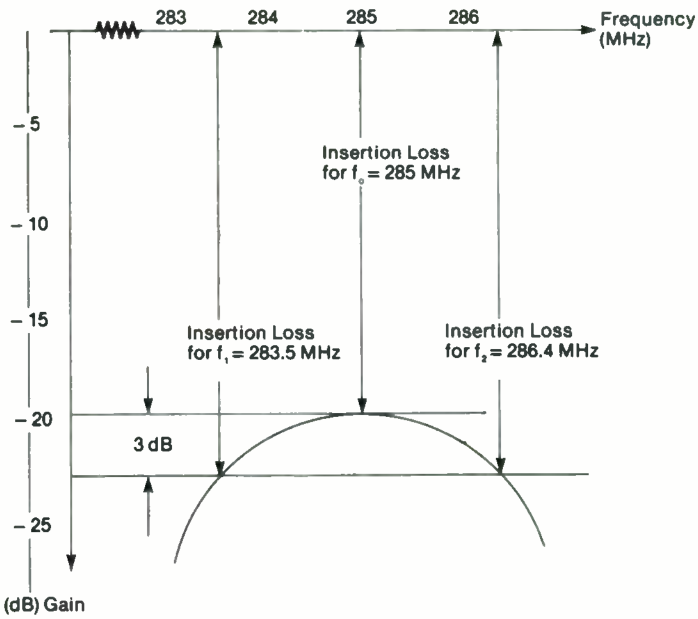


Figure 3. The 3 dB frequencies, which are well defined points, are used to calculate f_0, and Q.

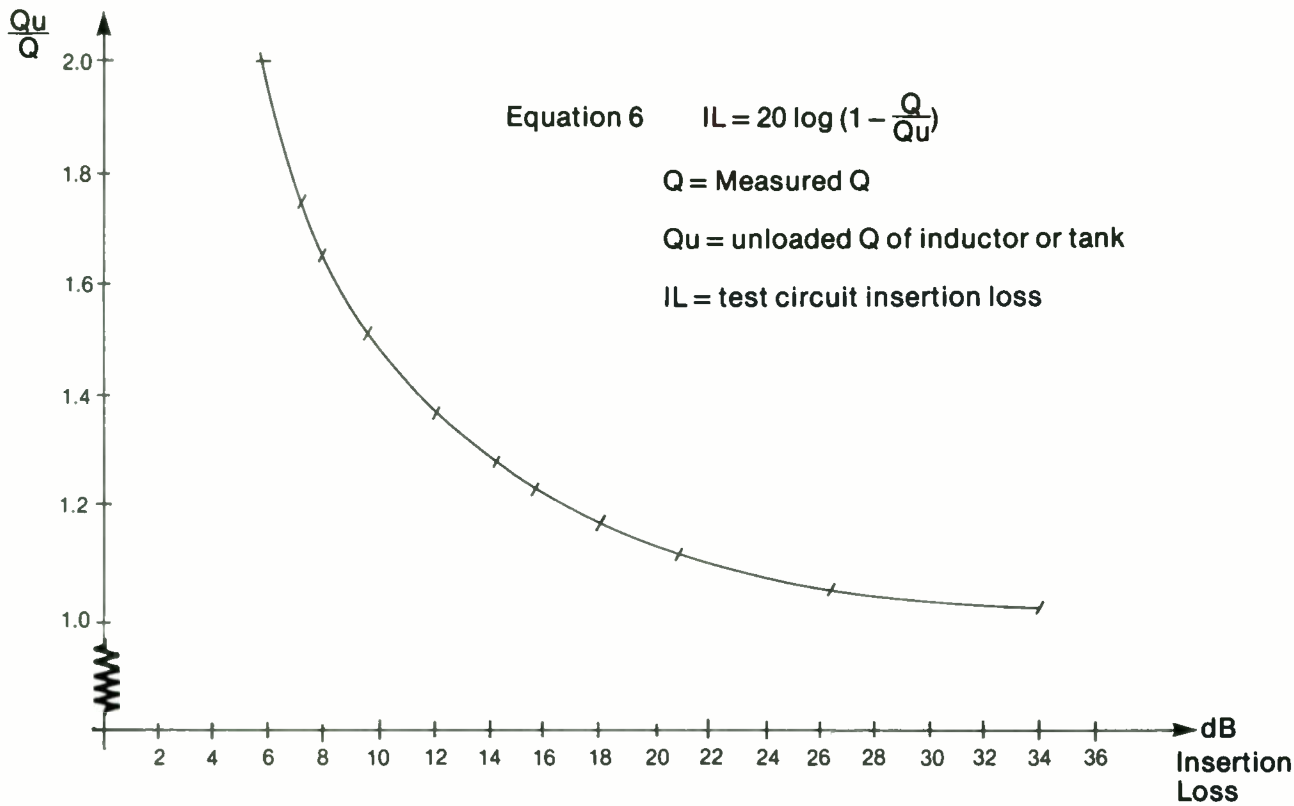


Figure 4. The insertion loss determines the ratio of the unloaded Q (Qu) to the measured Q.

L	X_L	R_{eff}	R_L	R_C	Q_{SP}	X_C	C_C	L
	$\omega_0 L$	QX_L	$Q_u X_L$	$R_{eff} = \frac{R_L 2R_C}{R_L + 2R_C}$	$\sqrt{\frac{R_C}{50} - 1}$	$\frac{R_C}{Q_{SP}}$	$\frac{1}{X_C \omega_0}$	$\frac{1}{\omega_0 (3.96 + C_C) \times 10^{-12} X_C}$
Nh	ohms	ohms	ohms	ohms		ohms	Pf	Nh
75	134.3	13,196	14,504	292,652	76.5	3825.5	.146	73.33
73.33	131.1	12,903	14,182	284,128	76.498	3714	.150	73.21

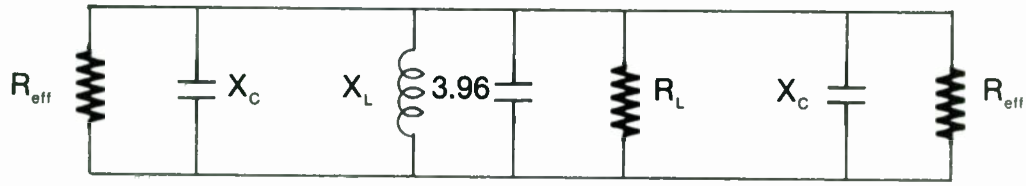


Table 1. The Insertion Loss value is used to resolve the inductance more accurately.

6) is independent of the coupling capacitors' values or coupling mechanism.

From Figure 4, observe that when the insertion loss is greater than 21dB, the ratio of Qu to Q is less than 1.1, or within ten percent error. The insertion loss value can be used to determine the calculated inductance error due to stray coupling capacitance in the measurement circuit, by using it to solve for Q and Qu, and then solving for the inductance values.

Although a closed form solution for the inductance error is possible, I suggest the following iterative approach (illustrated in Table 1).

1. Use the value of L to calculate X

$$X_L = \omega_0 L$$

2. Calculate the effective resistance of the circuit

$$R_{eff} = QX_L$$

3. Calculate the inductor resistance

$$R = Q_u X_L$$

4. Calculate the shunt equivalent of the load and source resistances. (R_C)

$$R_C = \frac{R_{eff} R_L}{2(R_L - R_{eff})} \text{ from } R_{eff} = \frac{(R)(2R_C)}{2R + R_L}$$

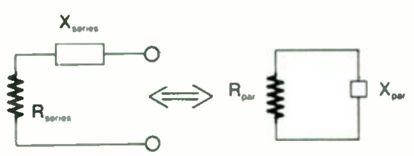
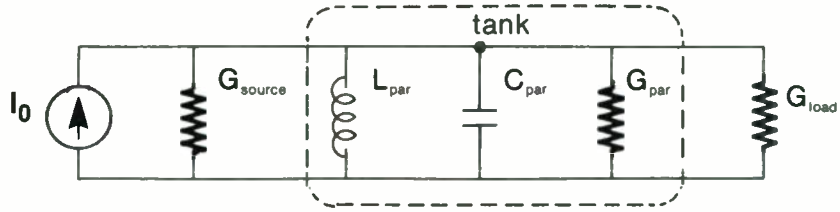


Figure 5. The parallel to serial transformation is used to establish the series equivalent circuit.



Power to load = V^2G_{load}
Power to Source = V^2G_{source}
Power to tank = V^2G_{par}

$X_{par} = X_L = X_C$

$Q_u = (\frac{1}{X_{par}})(\frac{1}{G_{par}})$

Figure 6. The Parallel equivalent of the test circuit is a filter whose midband IL relates the Q of the filter to the unload Q (Q_u) of the tank.

5. Calculate the Qsp of the series to parallel transformation

$$Qsp = \sqrt{R_C/50 - 1}$$

from Equation 2

6. Calculate $X_C = R_C/Q_{sp}$

7. Solve for the coupling capacitance C_C

$$C_C = 1/X_C\,\omega_0$$

8. Solve for $L = \frac{1}{2\pi(3.96 + C_C)\,X_C}$

$= 73.2$ uh

use the calculated value for L or recycle.

A measurement error in the resonant capacitor also affects the calculated inductance value. For example, an error of 0.1 pF in the capacitance measurement provides less than 1.73 nH uncertainty in the inductance value as shown by Equation 10 (Appendix IV).

$\Delta L = -1/(\omega_0 C)^2 = -1.72$ nH
where $\Delta C = 0.1$ Pf
$\omega_0 = 2\pi \times 285 \times 10^6$
$C = (3.96 + 0.15 + 0.15)$ pF
$= 4.26$ pF

Conclusion

A measurement technique to measure inductance and unloaded Q has been illustrated by an example. The measurement is at the frequency of operation and includes stray capacitance effects. The measurements use non-exotic laboratory equipment and relies on an accurate measurement of capacitance, insertion loss and frequency. The basis for the measurement is the construction of a narrowband filter whose Q approximately equals the unloaded Q of the inductor or tank being measured.

Appendix I.

Equations for series to Parallel Transformations.

The parallel equivalence of the series load resistance and coupling reactance is shown in Figure 5. The transformation equations are calculated by solving for the series equivalent of the shunt circuit.

$$Z_{series} = \left(\frac{R_{par}(jX_{par})}{R_{par} + (jX_{par})}\right)\left(\frac{R_{par} - jX_{par}}{R_{par} - jX_{par}}\right)$$

Multiplication of Z series to rationalize denominator

$$Z_{series} = \frac{R_{par}(X_{par})^2 + jX_{par}(R_{par})^2}{R^2_{par} + X^2_{par}} = R_{series} + JX_{series} \qquad (1)$$

Define Q (of the reactance)

$$Q = \frac{R_{par}}{X_{par}} \qquad (2)$$

then from Equation 1 and Equation 2

$$R_{series} = \frac{R_{par}}{1 + Q^2} \text{ or } R_{par} = R_{series}(Q^2 + 1) \qquad (3)$$

$$X_{series} = \frac{X_{par}Q^2}{Q^2 + 1} \text{ or } X_{par} = X_{series} = \frac{(Q^2 + 1)}{Q} \qquad (4)$$

for large Q, Q >> 1
$X_{par} = X_{series}$
please note from Equations 2, 3, and 4

$$Q = \frac{X_{series}}{R_{series}} = \frac{X_{par}Q^2}{R_{par}} = \frac{R_{par}}{X_{par}} \qquad (5)$$

Appendix II.

The series coupled source (and load) are transformed to their parallel equivalents by applications of the equations derived in Appendix 1.

$$X_{series} = \frac{1}{2\pi f_0 C} = 5.58 \times 10^3 \text{ ohm}$$

where $f_0 = 285$ MHz
$C = 0.1$ pF

Applying equation 5

$$Q = \frac{X_{series}}{R_{series}} = \frac{5.58 \times 10^3}{50} = 111.7$$

where $R_{series} = 50$ ohms

from Equation 3

$$R_{par} = R_{series}(Q^2 + 1) = 50\{(111.7)^2 + 1\} = 624{,}000 \text{ ohms}$$

from Equation 4

$$X_{par} = X_{series}\frac{(Q^2 + 1)}{Q^2} = (5.58 \times 10^3)\left(\frac{(111.7)^2 + 1}{(117.7)^2}\right) = 5.58 \times 10^3$$

for large Q, as we've shown
$X_{par} = X_{series}$
and
$C_{par} = C_{series} = 0.1$ pF

$$C_{resonant} = \frac{1}{\omega_0^2 L} = \frac{1}{((2\pi(285 \times 10^6))^2(75 \times 10^9)} = 4.16 \text{ pF}$$

$$C_{resonant} = Cpar_{source} + Cpar_{load} + C_R$$

Appendix III.

The insertion loss (IL) of the test circuit relates the filter Q to the Qu of the tank, or inductor, by Equation 6.

$$IL = 20 \log (1 - Q/Qu) \quad (6)$$

where Q = test circuit Q
Qu = tank or inductor Qu

The circuit is shown in Figure 6

$$P_{actual} = \text{Power delivered to load} = V^2G \text{ load}$$

$$= \frac{(I_0)^2 (G_{load})}{(G_{source} + G_{par} + G_{load})^2}$$

$$P_{ideal} = \text{Power delivered to load for } Qu = \infty \text{ or } G_{par} = 0 = \frac{(I_0)^2 (G_{load})}{(G_{source} + G_{load})^2}$$

$$\left(\frac{P_{actual}}{P_{ideal}}\right)^{1/2} = \frac{G_{source} + G_{load}}{G_{source} + G_{par} + G_{load}} \quad (7)$$

from Equation 2

$$Q = \frac{R_{total}}{X_{par}} = \left(\frac{1}{X_{par}}\right)\left(\frac{1}{G_{total}}\right)$$

or

$$G_{total} = \left(\frac{1}{Q}\right)\left(\frac{1}{X_{par}}\right)$$

for P_{actual} that is $0 < G_{par} < \infty$

$$G_{total} = G_{source} + G_{par} + G_{load} = 1/QX_{par} \quad (8)$$

$$G_{total} = G_{source} + \frac{1}{Qu\,X_{par}} + G_{load} = \frac{1}{QX_{par}}$$

$$\text{since } Qu = \left(\frac{1}{G_{par}}\right)\left(\frac{1}{X_{par}}\right)$$

$$G_{source} + G_{load} = \frac{1}{Q\,X_{par}} - \frac{1}{Qu\,X_{par}} \quad (9)$$

Substitution of Equation 8 and 9 into Equation 7

$$\left(\frac{P_{actual}}{P_{ideal}}\right)^{1/2} = \frac{\left(\frac{1}{Q\,X_{par}}\right) - \left(\frac{1}{Qu\,X_{par}}\right)}{\frac{1}{QX_{par}}} = 1 - \frac{Q}{Qu}$$

$$\text{Since insertion Loss (IL)} = 20_{\log} \left(\frac{P_{actual}}{P_{ideal}}\right)^{1/2}$$

$$\text{Then IL} = 20_{\log} \left(1 - \frac{Q}{Qu}\right)$$

As the Q of the test circuit approaches the unloaded Q (Qu) of the tank, the insertion loss (IL) increases as shown in Figure 4.

Appendix IV.

The inductance error due to capacitance measurement error is given by Equation 10.

$$\Delta L = \left(\frac{-1}{(\omega_0 C)^2}\right)\left(\Delta C\right) \qquad (10)$$

where ΔC = capacitance measurement error

$\omega_0 = 2\pi f_0$ (285 MHz)

C = total shunt capacitance

$= (3.96 + .03)$ pF $= 4.26$ pF

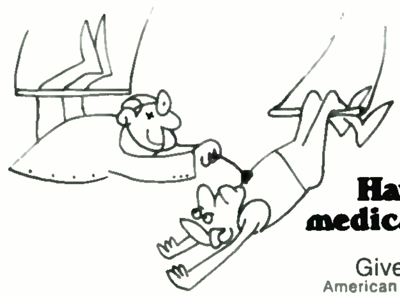
It is then derived by taking the derivative of the resonance equation.

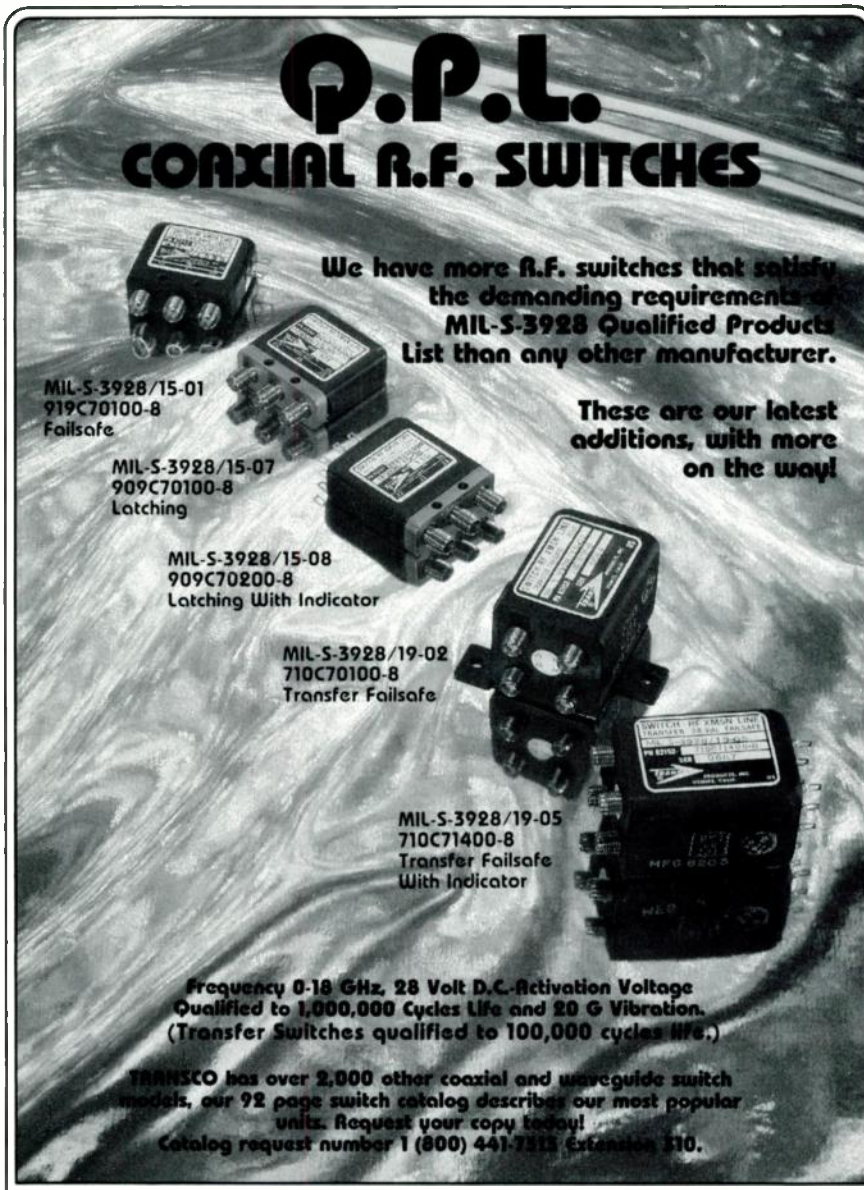
$$Lc = 1/\omega_0^2$$

$$dL \times C + L \times dC = 0$$

$$dL = \frac{L}{C} \times dC = -\frac{LC}{C^2} \times dC$$

$$dL = \left(\frac{-1}{(\omega_0 C)^2}\right) dC \qquad (11)$$





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A comparison with earlier VMOS technology specifications points up clearly the specifications of the Isofet that contribute to its superior performance. Power gain is typically 6 to 10 dB higher. The F_T is typically four times as high. What's more, the gate-to-drain capacitance (C_{RSS}) is typically 1.0 pf — or one-fourth that of other MOSFET devices. This accounts for its superior input/output isolation thereby easing stable amplifier design. There is less likelihood of spurious operation.

Isofet devices are also more rugged, can survive VSWR of infinity, and are also more tolerant insofar as safe operating region.

The Isofet technology employs a shorter channel, shorter base-width and narrower diffusion. Consequently, since mobility is some 30% higher in the surface region, the F_T is much higher than in many comparable MOSFET devices — as high as 5 GHz.

Another benefit is the dramatically reduced resistance of the parasitic bipolar transistor in parallel with the FET. It is typically one-sixth as large as the corresponding resistance in a VMOS design. This reduces the likelihood of secondary breakdown.

Isofet MOSFET technology also allows for a higher operating voltage. The lower ON resistance also renders the Isofet less prone to thermally-induced secondary breakdown since the safe operating area of the device is only limited by the breakdown voltage of the device and its thermal dissipation ability.

Acrian's Isofet technology departs markedly from VMOS technology by positioning the gate on the surface of the device, rather than in the "V-groove". Thus the channel is near the surface where mobility is higher, thereby raising the F_T. Also, the channel is much shorter, which accounts for the high F_T.

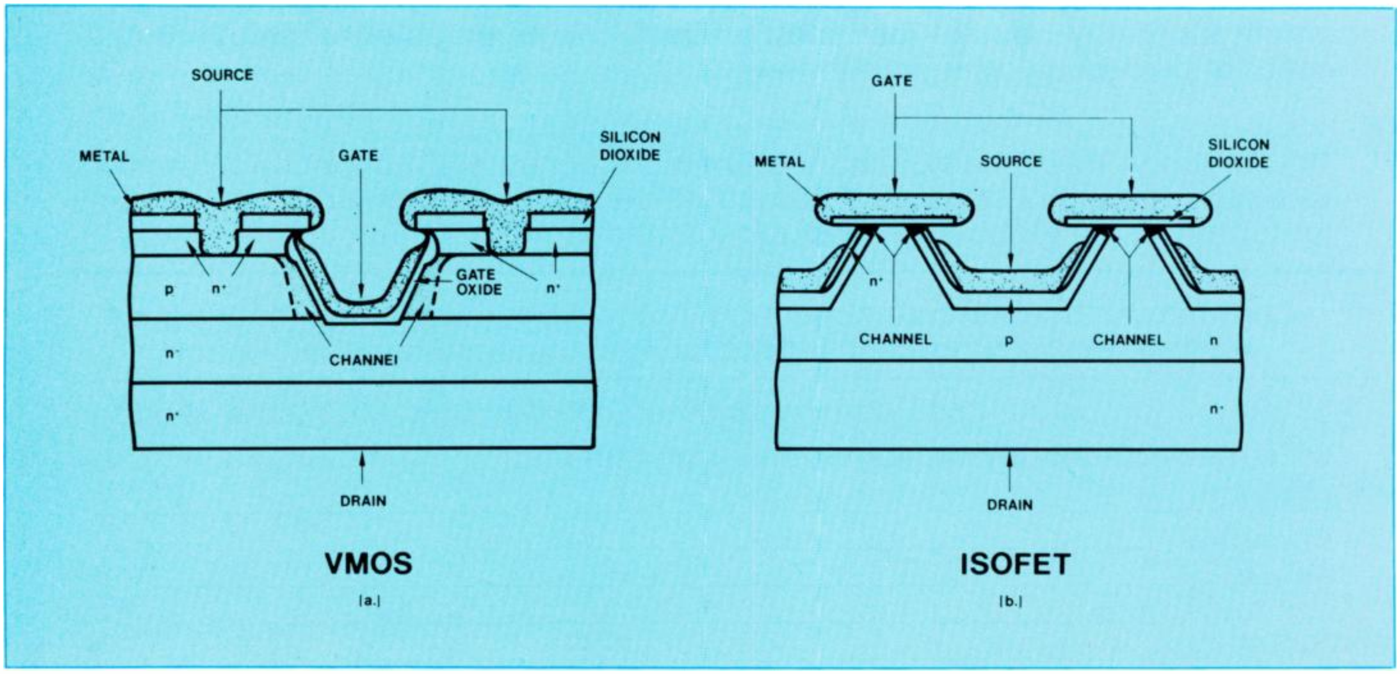
In the early 1970s Westinghouse developed and patented a shadow-isolated VMOS transistor. This approach entailed etching silicon from beneath a silicon-dioxide roof and then using the roof to shadow an evaporated metal that formed the gate. The Westinghouse process entailed elaborate fixturing to aim the evaporated metal properly at the channel under the roof.

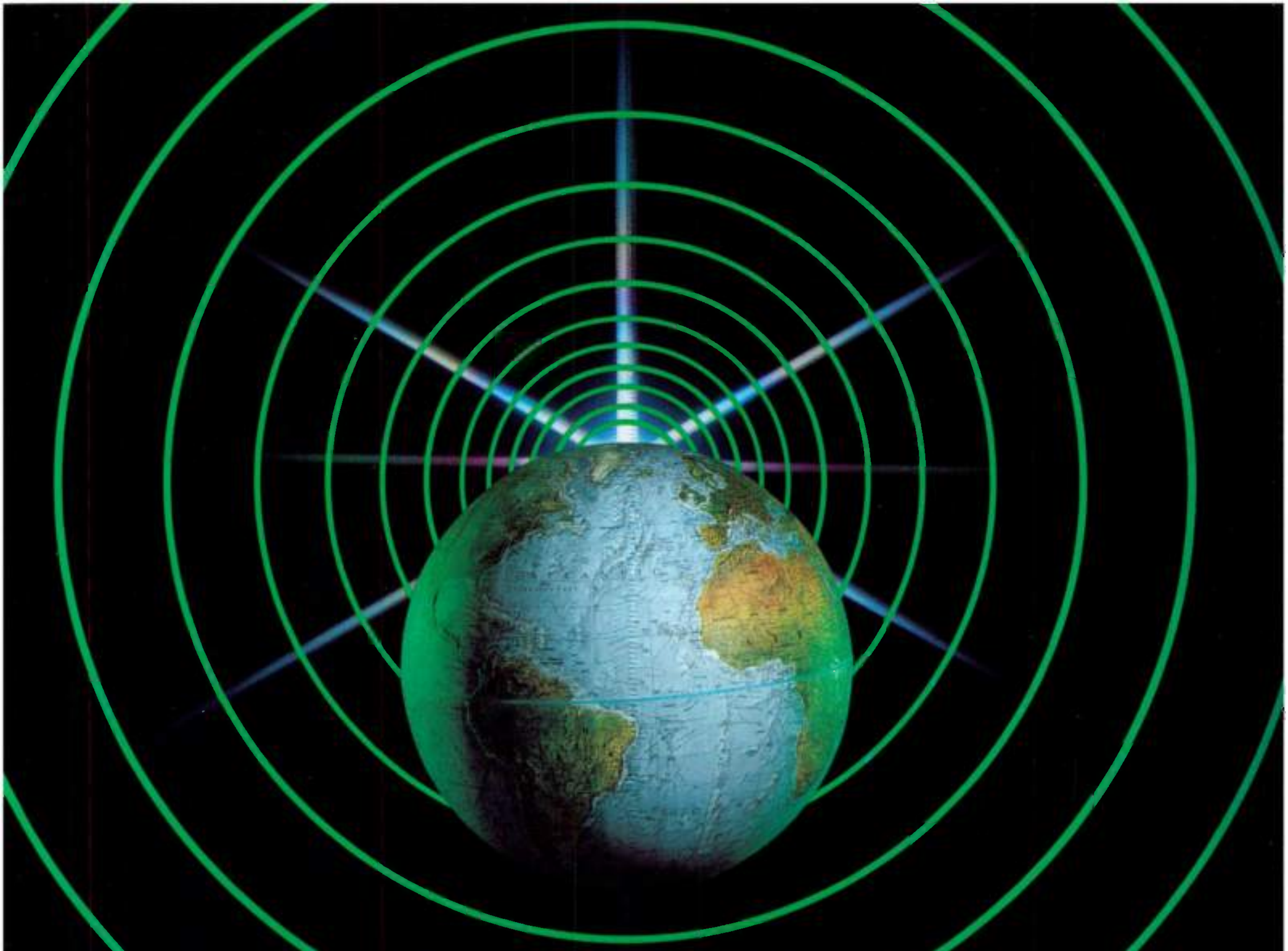
This process produced VMOS devices that operated in the 1.0 GHz region. However, this technology was never put into production.

In 1979 Siliconix introduced a VHF shadow-isolated VMOS transistor utilizing a similar oxide roof. However, their device was evaporated in a normal rotating planetary system, completely filling the bottom of the V-groove with metal.

Isofet™, Acrian's shadow DMOS device also utilizes the rotating planetary evaporation technique. However, the Isofet technology differs in that the channel is raised to the surface of the device thereby exploiting the higher surface mobility. What's more, since the channel is parallel to the surface, the channel length is shorter and capacitance is lower.

Isofet is MOSFET technology or metal gate technology. Whereas silicon gate technology is confined to 5.0 MHz, metal gate technology is less lossy than polysilicon. The shadow-isolated overhang acts as a self-aligning mask. Alignment is 'triply' self-aligning. Thus Isofet MOSFETs are cheaper to make since they require fewer masking operations. □





40-Watt UHF Transistor

Acrian's new UMIL40FT 40-watt, UHF broadband power transistor enables designers to extend the inherent advantages of the MOSFET in wideband, high-power amplifiers through 500 MHz. Because this device utilizes Acrian's proprietary Isofet™ technology, amplifier designs can be configured which exhibit higher stage gain, enhanced input/output isolation and better thermal stability than is possible with other MOSFET devices.

The UMIL40FT consists of two Isofet devices internally connected in push-pull. It exhibits 10-dB more gain than a bipolar device and 3-dB more gain than MOSFET devices previously available for r-f designs.

The new Acrian Isofet is well suited for AM and FM aircraft/land communication, 2-30 MHz SSB, 30-88 MHz programs, Military AM/FM in the 100-400 MHz region and broadband instrumentation and countermeasures in the 200 to 500 MHz range.

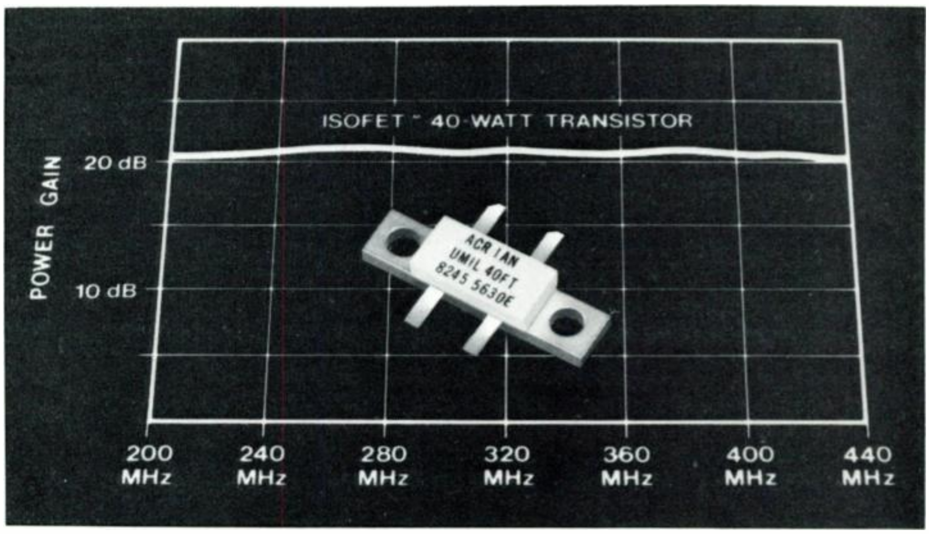
One of the characteristics of the UMIL40FT which appeals to power amplifier designers is that a virtually-flat, power-output profile is readily achieved without the need for gain sloping. This technique has frequently been required in bipolar circuits intended to operate over octave bandwidths. Sophisticated gain sloping networks using R, L and C or quadrature combiners are required with bipolars.

Another feature which simplifies design is the fact that the input impedance can be adjusted to 12.5, 25 or 50 ohms. This eliminates any requirement for complex input matching networks. Simple input transformers are all that are necessary.

Biasing can be accomplished simply with a resistor. Sensitive tracking diodes with feedback are not required. Also the Acrian UMIL40FT Isofet is more forgiving in regard to VSWR, has a large safe operating area, and has a maximum dissipation of 100 watts.

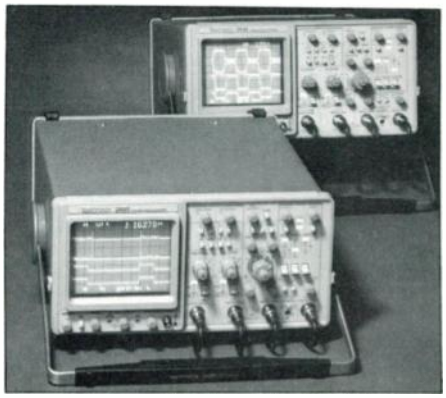
FETs have the additional benefit that they can be high-level modulated through simple voltage control of the gate. Thus, simple logic circuits can be utilized to directly modulate the UMIL40FT. There is no need to simultaneously modulate the voltage and drive with the attendant complexities involved.

Price of the UMIL40FT is $105.00. Delivery is 3-4 weeks. **Acrian, Inc., Cupertino, CA 95014, or please circle INFO/CARD #92.**



2400 Series Oscilloscopes

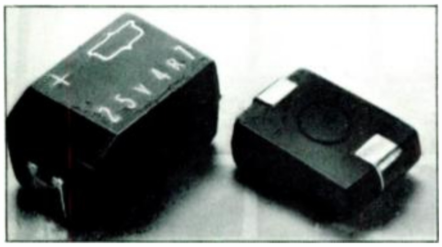
Using today's most advanced technologies, Tektronix 2400 Series portable oscilloscopes establish a new industry standard. Two 2400 scopes are being introduced, the 2445 and 2465. The 2445 and 2465, analog scopes with significant benefits for digital applications, are designed to replace the Tektronix 465B and 475. At $3,140 and $4,600, the new scopes' broader bandwidths, four channels, extensive CRT readouts, faster sweep speeds and increased timing accuracy set the standard for scopes to come. The 2465 has 300 MHz bandwidth, 500 picosecond per division sweep speed, and an auto-level trigger feature with trigger bandwidth greater than 500 MHz. The 2445 has 150 MHz bandwidth, 1 nanosecond per division sweep speed, and the same triggering features with trigger bandwidth to 250 MHz. The 20.3-pound scopes have four full-bandwidth vertical channels, each with calibrated scale factors and positioning controls. Another feature



is the channel 1 and 2 propagation delay matching adjustment. **Tektronix, Inc., Beaverton, OR 97077 or PLEASE CIRCLE INFO/CARD #140.**

Leadless Aluminum Electrolytic Chip Capacitors

United Chemi-Con, Inc. announces the world's first aluminum electrolytic chip capacitor. The ALCHIP-S is ideal for hybrid circuit and leadless printed circuit board applications. This series is available with a capacitance range of 0.1 to 22 uF ± 20%, in six voltages from 6.3 to 50 VDC. Mechanically, the ALCHIP-S utilizes a molded plastic case capable of withstanding 230°C



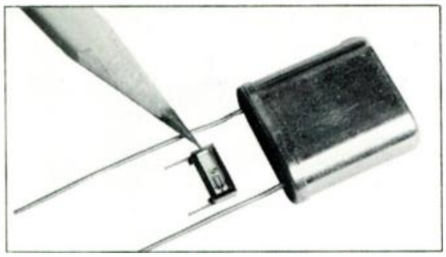
for 10 seconds, allowing assembly by standard re-flow soldering techniques. Price: $.07 each in 1,000 quantities. **United Chemi-Con, Inc., Rosemont, IL 60018, INFO/CARD #139.**

SPDT Module Switches With Driver

M/A-COM Microwave Circuits, Inc. Hudson, NH announces the MA-8345 series of SPDT module switches with integrated drivers. These hermetically constructed, connectorless units cover the frequency range of .5 to 18 GHz and offer a switching speed of 10 ns. They also offer high isolation and TTL compatibility. The MA-8345 series will find use in RF switching modulation and channel selection. Designed specifically for broadband ECM requirements, their light weight and small size make them an excellent choice for airborne and missile applications. **M/A-COM Microwave Circuits, Inc., Hudson, NH 03051, or INFO/CARD #137.**

Ultra-Miniature Quartz Crystals

Ultra-miniature quartz crystals that operate at frequencies up to 1.5 MHz have been introduced by Statek Corp. The manufacturer claims that they are the smallest crystals available at any frequency in the range 10 kHz to 1.5 MHz. They are 48 times smaller in volume than the industry-standard HC-33 crystal package and occupy only one-tenth as much board space. This expanded CX-1 Series is available in a hermetically-sealed ceramic package



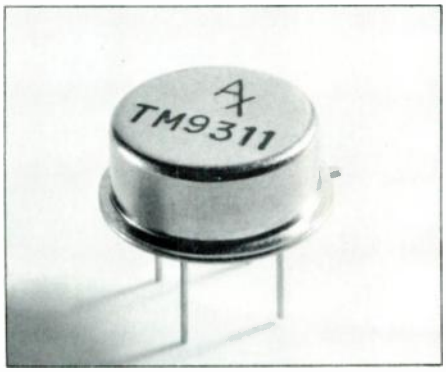
with 0.300-inch lead spacing and in a leadless version for surface mounting. Prices: Less than $10.00 each in 100 quantity. Delivery: Twenty-three standard frequencies available from stock. **Statek Corp., Orange, CA 92668, or INFO/CARD #138.**

Plotter Module Adds Graphics To HP-41 Handheld Computers

A low-priced system for generating graphics and bar code with the HP-41 handheld computers is now available from Hewlett-Packard Company. With the introduction of the plotter module at a $75 list price, HP-41 owners can link their handheld computers to the HP 7470 color graphics plotter and plot charts, graphs and bar code. The HP-41 handheld computer, with the plotter module, is connected to the HP 7470 graphics plotter via the Hewlett-Packard Interface Loop (HP-IL). A new HP-IL-compatible version of the plotter (HP 7470 Option 003) is being introduced concurrently with the plotter module. The new module plugs into the HP-41 and adds a set of commands and functions to the handheld computer that enable the user to plot charts, graphs and bar code. **Call your local Hewlett Packard Sales Office or INFO/CARD #122.**

RF Amplifier

Amplifonix has announced a new RF Hybrid Amplifier, Model TM9311 in a TO-8 hermetic package. This new design provides a gain of 16 dB typical over the frequency range of 5 to 1000 MHz. Other specifications include: Noise Figure, 2.4 dB typ; Power Out @ 1 dB comp., −1 dBm; Max. VSWR (In/Out); 2:1; Current @ 15V, 9 ma.; Temperature range, −55 to +100°C.



All units meet MIL-STD-883B screening. Price for 1-9 is $90; delivery from stock. **Amplifonix, Inc., Bristol, PA 19007, or INFO/CARD #135.**

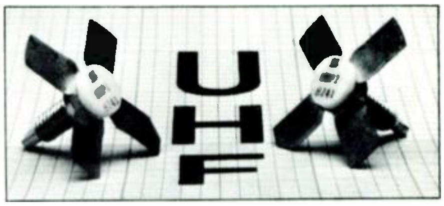
Micro-Miniature Band Pass Filters

RLC Electronics, Inc. announces the introduction of Micro-Miniature Band Pass Filters. These units offer excellent frequency response characteristics with low insertion loss. The small size capability is provided by utilization of miniaturized High Q devices in a microstrip mode. Standard units utilize low ripple, Chebychev design. Other re-

sponses available when desired. Filters are available with spurious free response to 12.4 GHz when required. Prices start at $250.00 in unit quantity. Delivery of small quantity in 6 weeks. **RLC Electronics, Inc., Mt. Kisco, NY 10549, or INFO/CARD #134.**

UHF Power MOSFETS

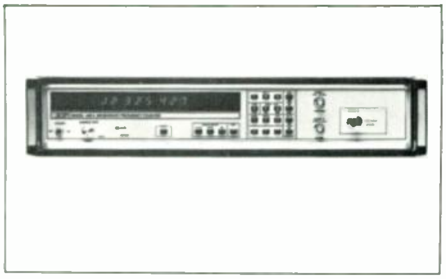
Siliconix has announced the immediate availability of the first commercially available 5 and 10 watt UHF



power MOSFETs. Designated the UMP-1 and UMP-2, the n-channel, enhancement-mode devices can deliver 5 watts and 10 watts, respectively, while operating at frequencies up to 400 MHz. Specified minimum power gains are 10 dB for the UMP-1 and 7 dB for the UMP-2. The introductions of the new UHF parts extends Siliconix MOSPOWER devices beyond the HF/VHF operating frequency range into the UHF region. **Siliconix Inc., Santa Clara, CA 95054, or INFO/CARD #133.**

Counter For Communications Use

EIP Microwave has announced a new configuration of their 10Hz to 18GHz 545 Microwave Frequency Counter designed especially for the communications industry. The manufacturer claims that this new unit, designated the 545-W29, offers several



advantages to telco transmission engineers and craftsmen in the alignment, repair and verification of microwave communications equipment. These advantages include: 1) An input YIG filter for frequency selective power measurements. This allows output power to be measured right at the feedhorn test-point for quicker and easier adjustment of transmitter power levels on multiple carrier systems. 2) Measurement of final power and frequency, without turning off modulation. 3) A build-in power meter for direct power readings. 4) Five watts of burn-out protection to prevent counter failure due to accidental overload. Priced at $7400, the 545-W29 weighs only 20 lbs. **EIP Microwave, Inc., San Jose, CA 95134, or please circle INFO/CARD #131.**

Sampling Oscilloscope System

I.T.I. S-500 is a dual trace delayed Equivalent-Time Sampler used with any oscilloscope capable of X-Y display to form a Sampling Oscilloscope system. It provides an input bandwidth of dc-700MHz and a Time Base from 500ps/div. to 100ns/div. A P-700 FET-Probe with dc-500MHz bandwidth can be used to acquire signals from high impedance sources to the sampler.

The S-500 and the P-700 are priced at $480 and $63 respectively. **International Test Instruments Corp., Cupertino, CA 95015, or please circle INFO/CARD #130.**

Leadless Inductive Components

An expanded line of standard and miniaturized West-Chip™ leadless ceramic chip inductors is offered by West-Cap Arizona, a subsidiary of SFE Technologies, San Fernando, California. They are designed for use in thick film, high-density hybrid microcircuitry, resonant circuits and decoupling applications, as well as in low power signal applications. Mounting is by either solder reflow or conductive epoxy. Core material is ferrite, powdered iron or phenolic, depending upon the inductive value of the device. Standard WCS series inductors, mounted on .125" × .150" ceramic substrate, are supplied in a standard range of .010 uH to 1000 uH. Custom devices are available up to 47 mH. **West-Cap Arizona, San Fernando, CA 91341 or INFO/CARD #129.**

SMA Coaxial Attenuators

SMA Coaxial Attenuators are now available with a frequency range from D.C. to 26.5 GHz. Precision machining and minimum internal discontinuities eliminate moding problems normally associated with standard SMA connectors in the 18 to 26.5 GHz band. The connectors will mate non-destructively and provide optimum performance with APC-3.5 connectors. Attenuation values from 1 to 30 dB are available in two subminiature sizes. Typical frequency variation is less than ± 0.5 dB from nominal and VSWR is less than 1.50 up to 26.5 GHz. **KDI Pyrofilm Corporation, Whippany, NJ 07981, or INFO/CARD #128.**

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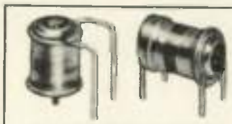

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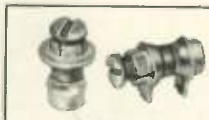

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RF Plug-Ins HP Sweeper

Hewlett-Packard has expanded the range and performance of its microprocessor-based sweep oscillator family (HP 8350) with the introduction of five new RF plug-ins. The units, which provide new frequency coverages and/or new performance capabilities, are:

U.S. List Prices and Delivery

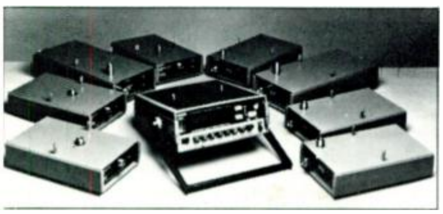
Model No.	U.S. List Price	Delivery
HP 83572A (26.5—40 GHz)	$14,540	16 weeks ARO
Option 001-external leveling add	1,605	
HP 83592B (0.01—20 GHz)	26,580	12 weeks ARO
HP 83525B (0.01—8.4 GHz)	15,540	4 Weeks ARO
HP 83540B (2.0—8.4 GHz)	10,280	4 Weeks ARO
HP 86251A (7.5—18.6 GHz)	10,780	12 weeks ARO
HP 11869A adapter (for use with with HP 8350)	405	2 weeks ARO
HP 8350A mainframe	4,565	2 weeks ARO

Call local Hewlett-Packard sales office or INFO/CARD #132.

Microwave Multimeter Active Heads

New Microwave Units are now available for the Microwave Multimeter System.

MODEL	FREQUENCY
7123	1.7—2.3 GHz
7123-S1	2.2—2.7 GHz
7144	6.5—7.2 GHz



With the addition of these Communications and Radar Band Units, the active heads cover all the significant bands in the frequency range from 1.7 to 12.7 GHz. **The Narda Microwave Corporation, Hauppauge, N.Y. 11788 or INFO/CARD #126.**

Bench Test Receiver 9 kHz to 30 MHz

The programmable Test Receiver ESH3 is used for measurement and demodulation of Am, DSB, SSB, pulse-modulated and FM signals as well as interfering sinewave and pulse signals in the range 9 kHz to 30 MHz. High overdrive capacity, excellent dynamic range, elaborate measurement/evaluation capabilities and a large number

of extras make the ESH3 useful for all applications concerned with radio detection and EMC measurements. The measurement range of −30 to +137 dB (V) makes the ESH3 Test Receiver useful as an automatic, highly accurate, selective voltmeter for laboratories, test departments and inservicing without requiring additional accessories. **Polarad Electronics, Inc., Lake Success, N.Y. 11042, or please circle INFO/CARD #127.**

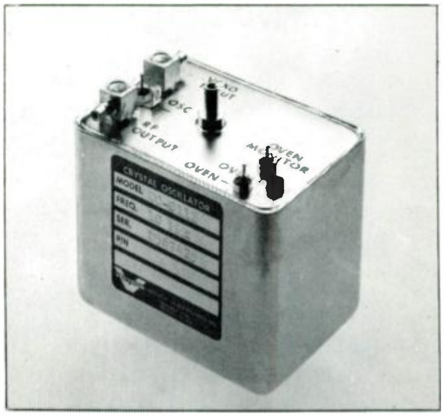
100 Hz To 1000 MHz Noise Generator

Micronetics has combined the frequency range of 7 different PNG5100 models into one in its new PNG5200 Programmable Noise Generator to cover 10Hz to 1000 MHz frequency band with an output of +10 DBM (10 mw). A computer, operating over the IEEE-488 bus, can control the output level over a range of 99dB in 1dB steps. The source can also be bus-controlled to "on" or "standby" states as is required for system noise-figure measurements. The instrument contains a Solid State Noise Source

that produces Gaussian noise with a 5:1 peak-to-RMS ratio, a wideband amplifier, a bandpass filter, a programmable attenuator, and an IEEE-488 bus interface. The requirements of these circuits are stringent, since they must result in the instrument's specified output flatness of ±2.5 dB over the frequency range. **Micronetics, Inc., Norwood, NJ 07648 or INFO/CARD #125.**

Replacement For Discontinued HP 10811B Crystal Oscillator

Model CO-811B offers mechanical and electrical interface interchangeability with the recently discontinued Hewlett Packard Model HP10811B High Stability 10 MHz Crystal Oscillator. It uses an SC-cut crystal for fast re-stabilization and meets the HP10811B noise floor of −160dBc/Hz, aging of 5×10^{-10} per day, and dual temperature stabilities of 2.5×10^{-9} total over 0°C to +71°C and 4.5×10^{-9} total over −55°C to +71°C. A series of less stringent stability models is also offered in the same mechani-



cal package with associated cost savings. **Vectron Laboratories, Inc., Norwalk, CT 06850 or INFO/CARD #124.**

"T" Attenuator Module

Alpha Industries, Inc., announces the availability of a new "T" attenuator module (#DAA 5125) that features monolithic array construction with multiple beam lead diode junctions. The device utilizes nitride and oxide passivations for high reliability and is supplied in a variety of resin encapsulated packages, or in chip form, for use in microstrip or stripline applications. Operating temperature is −55°C to +150°C, with a power handling of 250 MW. These modules are suitable for switching as well as attenuating applications. **Alpha Industries, Inc., Woburn, MA 01801 or please circle INFO/CARD #123.**

RF Wattmeter

A new THRULINE® Directional Wattmeter expands the usual single full-scale power level of its plug-in element to seven overlapping power ranges. Tailored for design or field-

service of CW and FM systems from 200 kilohertz (kHz) to 1000 Megahertz (MHz) and 1/4 watt to 10,000 watts, the new precision instrument uses special elements providing seven levels instead of one covering either 1/3/10/30/100/300/1000 watts or 10/30/100/300/1000/3000/10000 watts with ±5% accuracy OF READING over a full 37dB power range. The desired range is instantly selectable by a front-panel rotary switch which also includes a battery-level position. Elements are simply rotated for either forward or reflected power measurement. Price of this unique Bird RF Wattmeter is $495, 7-in-1 Elements are $125-175. **Bird Electronic Corporation, Cleveland (Solon), Ohio 44139, or please circle INFO/CARD #120.**

DC-18 GHz Multi-Position Coaxial Switches

Teledyne Microwave, announces its new line of CS-38 broadband subminiature coaxial switches with low power Schottky TTL circuits with decoders. The control input is a 3-bit parallel word that is decoded internally to select one of the 3 to 6 available positions. RF performance at 18 GHz is 1.5:1 VSWR, 0.5 dB insertion loss, and isolation of 60 dB. Options available include indicators, actuator voltages of 12, 15, 20, and 28 VDC, and power connectors. **Teledyne Microwave, Mountain View, California 94043, or INFO/CARD #118.**

Application Notes

RF Hermetic Package Mounting Techniques

An application note on techniques for mounting TRW RF linear amplifiers in hermetic packages is available from TRW Semiconductors. The 4-page note describes guidelines for obtaining a suitable interface between a printed circuit board and either a hermetic single-in-line package or a TO-8 package. High reliability screening capabilities for military applications are also outlined. **TRW Semiconductors, Lawndale, CA 90260, or please circle INFO/CARD #110.**

Transistor Application Note

California Eastern Laboratories, Inc., offers NEC's Technical Note TN82401 titled "Low Distortion Microwave Linear Power Transistor." This paper described the design and experimental results for high gain and low distortion transistors used in a CATV main cable repeater at up to 500 MHz. Diagrams are located throughout this informative paper. **California Eastern Laboratories, Inc., Santa Clara, CA 95050, or please circle INFO/CARD #113.**

OP Amp Application Note

Bill Olschewski, author of many articles on analog circuit design, has written a comprehensive applications note on power op amps covering such subjects as safe operating area, inductive loads, emf generating loads, bridge circuits, speed and position control circuits, cathode ray drive circuits and programmable power supplies. **Apex Microtechnology Corp., Tucson, AZ 85714 or INFO/CARD #112.**

Auto-Tracking Filter

A new 8 page applications brochure describes Integra's TMF-1800H (2-18.5 GHz) Tunable Microwave Filter. This note illustrates how the Filter can Auto-Frequency Track a signal source or be independently tuned under 488-Bus Control. With this filter, the harmonic output level of Sweepers or Synthesizers can be reduced to a −55 or −70 dBc level. The Tunable Filter can also Offset frequency track a signal source, making it extremely useful for test system applications. **Integra Microwave, Santa Clara, CA 95051, or INFO/CARD #111.**

New Literature

Attenuators Brochure

Weinschel Engineering has published a new document entitled EVERYTHING YOU WANTED TO KNOW ABOUT MIL-A-3933D ATTENUATORS BUT WERE AFRAID TO ASK. This informative synopsis presents all the aspects of specifying attenuators in accordance with MIL-A-3933. It answers such questions as, "What is the impact to the user?", defining CLASS I, II, III and IV. This 10 page document details the requirements invoked when adherence to MIL-A-3933 is requested including: First Article Inspections, Screening Inspections, Group A Inspection, Group B Inspection, Procurement Examples and Test Flow Charts for CLASS I, II, III and IV Attenuator Inspection. As an added extra, a complimentary copy of MIL-A-3933D with all current slash sheets will accompany this informative document. **Weinschel Engineering, Gaithersburg, MD 20877 or please circle INFO/CARD #109.**

Product Catalog

Kay Elemetrics Corp. new 20 page catalog describes its full line of attenuators, switches and R.F. Test equipment. The Attenuator product line includes Miniature Step, Standard Step, Rotary, Continuously Variable, Audio and OEM Programmable attenuators. Also described are the R.F. Mega switches, Pulsed Carrier Generators and Sweep Frequency Generator. **Kay Elemetrics Corp., Pine Brook, N.J. 07058, or INFO/CARD #108.**

Communications Antenna Catalog

A new catalog of commercial, industrial and military HF, VHF and UHF communications antennas and antenna systems is available from Hy-Gain, a division of Telex Communications, Inc. Specifications are stated in both the English and metric systems for antennas ranging from tiny quarter-wave whips for mobile applications to giant arrays requiring multiple acre installation for worldwide communications. When applicable, the U.S. military nomenclature and national stock numbers are included. **Hy-Gain, Lincoln, NE 68505 or please circle INFO/CARD #107.**

SIPMOS Catalog

The new 1982/83 SIPMOS Small Signal Transistor Catalog has been released by Siemens Components, Inc. The 30-page guide contains sections on SIPMOS technology, maximum ratings, characteristics, bi-polar comparisons, data sheets on the eight configurations and an interpretation of technical data. The transistors presented in the catalog are Siemens' initial offering of small signal SIPMOS FETs. Low drive requirements and fast switching speeds make them ideally suited to interface directly with logic circuits and microprocessors for control of motors, switching power supplies and drives for high powered transistors. **Siemens Components, Inc., Iselin, NJ 08830, or INFO/CARD #106.**

RF Coaxial Connector Catalog

AVA's new 82-5 catalog features F, UHF, BNC and Twinax connectors and the new Computer Switches. Crimp-type connectors are highlighted in the 16 page, two-color catalog. UG type connectors are cross-referenced in the handy index. Adapters for a variety of connector types are also shown in the revolutionary-for-AVA catalog. **AVA Electronics Corp., Drexel Hill, PA 19026 or INFO/CARD #105.**

Cable Product Catalog

A new catalog giving complete data on local area network cables, computer cables, and fiber optic cables is now available from the RF Cable Products Division of Times Fiber Communications, Inc., an affiliate of Insilco Corporation. The 24-page illustrated catalog features a four page selection guide and application notes on coaxial cables which engineers and purchasing agents will find most helpful. **Times Fiber Communications, Inc., Wallingford, CT 06492, or please circle INFO/CARD #104.**

YIG Devices Catalog

A new 48-page YIG Devices catalog is available from Watkins-Johnson Company. The catalog includes the company's full line of YIG-tuned oscillators, filters, integrated devices and harmonic generators. In addition to product listings, the catalog offers a technical introduction to YIG devices, selection guides and application notes. **Watkins-Johnson Company, Palo Alto, CA 94304, or INFO/CARD #103.**

Solid State Products Catalog

An eight page catalog describing Addington solid state products is available from Eaton Corporation's Communications Products Division. Among the products described are the Addington line of high reliability, thin film VCO's, mixer up/down converter subsystems, and RF cable and connector products. In addition, the brochure describes the division's advanced capabilities, equipment and quality assurance programs. **Eaton Corp., Sunnyvale, CA 94086, or please circle INFO/CARD #102.**

Video Delay Lines & Filters Catalog

Allen Avionics has issued a new 8 page catalog, No. 16V, encompassing the firm's complete line of Video and Pulse Delay Lines & Video Filters. Introduced is Hum Eliminator HEC1000, which eliminates hum and other interference in video lines caused by differences in ground potential. Also presented are Random Noise Measurement Networks and Pre-emphasis and De-Emphasis Wave Shaping Networks. A new Delay Equalized Lowpass Filter VSL4P5, is also introduced. **Allen Avionics, Inc., Mineola, NY 11501, or INFO/CARD #101.**

Thermistor Line Catalog

Designers planning to use negative temperature coefficient thermistors will find a wealth of useful information condensed in a new "pocket-size" catalog available from Dale Electronics. The "catalog" details the products of the Dale Western Thermistor Division. Specifications are given for most frequently specified styles, many of which are available for applications where interchangeability is required. Drawings are provided showing "typical" sensor assemblies which are available from Dale. Included in the "pocket catalog" are RT conversion tables and background information on thermistor applications. **Dale Electronics, Inc., Columbus, NE 68601, or INFO/CARD #100.**

Test Instruments Brochure

A new brochure describing telecommunications test instruments for digital radio, fiber optics, satellite communications systems and digital modems is available from Scientific-Atlanta, Inc. **Scientific-Atlanta, Inc., Atlanta, GA 30348, or INFO/CARD #99.**

Capacitor Catalog

A recently revised catalog with all of the newest capacitor offerings from TRW Capacitors is now available. This catalog contains all the information needed to select the appropriate type of capacitor for typical applications. Included are sections on the following types of wound capacitors: metallized polycarbonate dielectric, metallized polypropylene, metallized polyester, polyester dielectric, polystyrene dielectric, and polyester dielectric RC network. **TRW Capacitors, Ogallala, NE 69153, or please circle INFO/CARD #97.** □

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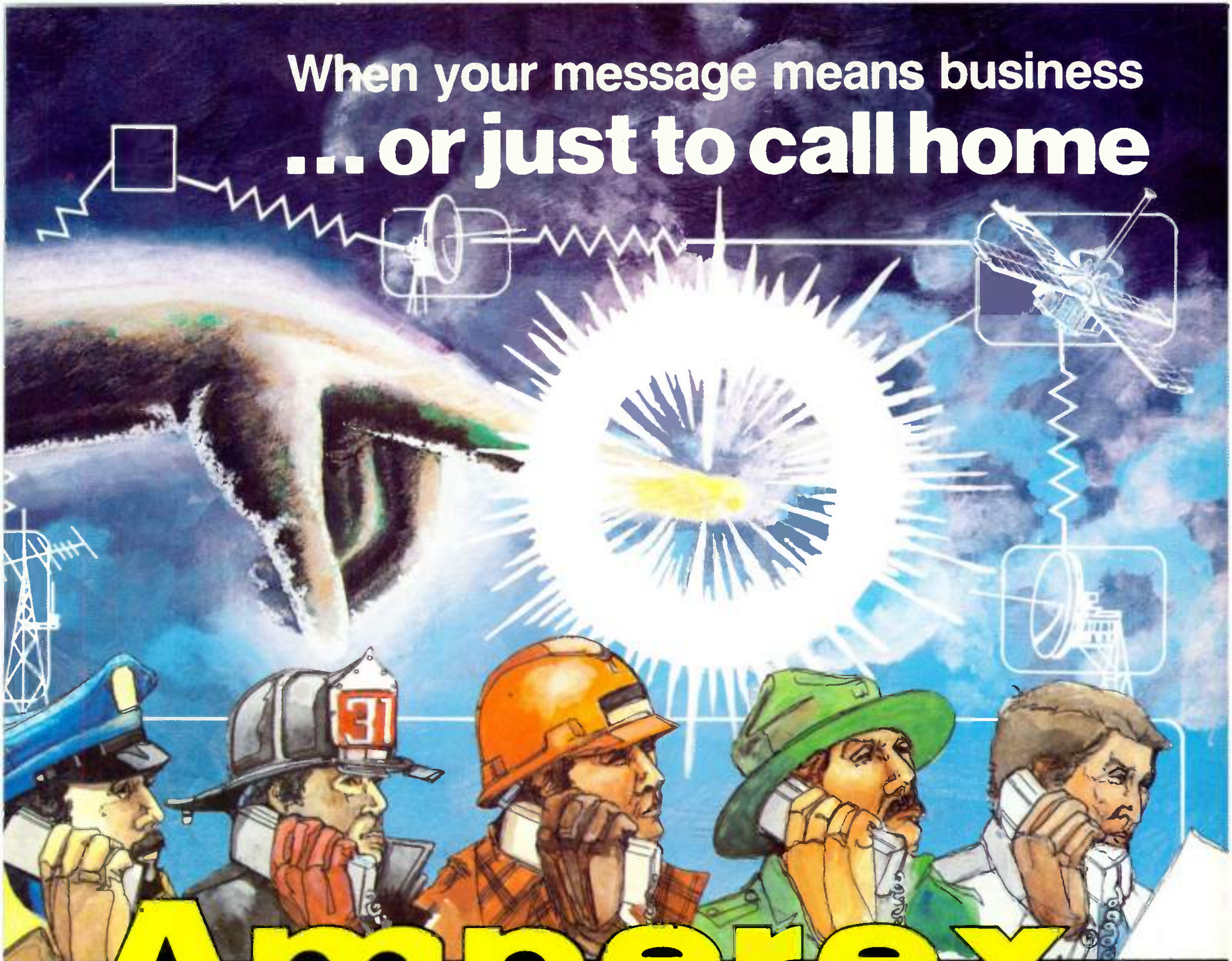

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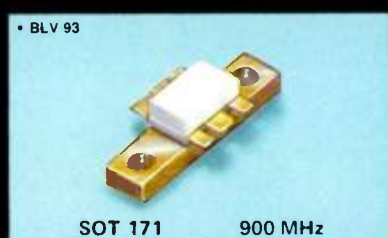
APPLICATION: 900 MHz Radio Telephone
POWER: 0.5 Watts
GAIN: ≥ 9.0 dB
FREQUENCY: 900 MHz
V_{CE}: 12.5 Volts
EFFICIENCY: ≥ 50%

INFO/CARD 1

APPLICATION: 900 MHz Radio Telephone
POWER: 2 Watts
GAIN: ≥ 6.5 dB
FREQUENCY: 900 MHz
V_{CE}: 12.5 Volts
EFFICIENCY: ≥ 50%

INFO/CARD 2



APPLICATION: 900 MHz Radio Telephone
POWER: 8 Watts
GAIN: ≥ 6 dB
FREQUENCY: 900 MHz
V_{CE}: 12.5 Volts
EFFICIENCY: ≥ 50%

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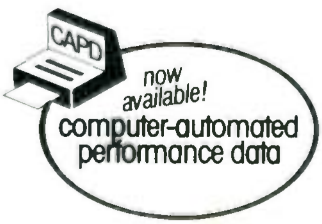
Noise Figure	7.5dB
Supply	+15V @ 690mA
Third Order Intercept	38 dBm Min.
Second Order Intercept	48 dBm Min.
Size	7" × 3¼" × 2⅛"h



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